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# MILITARY HANDBOOK

## GROUNDING, BONDING, AND SHIELDING FOR ELECTRONIC EQUIPMENTS AND FACILITIES

VOLUME I OF 2 VOLUMES  
BASIC THEORY

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GROUNDING, BONDING, AND SHIELDING FOR ELECTRONIC EQUIPMENTS AND FACILITIES

1. This standardization handbook was developed by the Department of Defense in accordance with established procedure.

2. This publication was approved on 29 December 1987 for printing and inclusion in the military standardization handbook series. Vertical lines and asterisks are not used in this revision to identify changes with respect to the previous issue due to the extensiveness of the changes.

3. This document provides basic and application information on grounding, bonding, and shielding practices recommended for electronic equipment. It will provide valuable information and guidance to personnel concerned with the preparation of specifications and the procurement of electrical and electronic equipment for the Defense Communications System. The handbook is not intended to be referenced in purchase specifications except for informational purposes, nor shall it supersede any specification requirements.

4. Every effort has been made to reflect the latest information on the interrelation of considerations of electrochemistry, metallurgy, electromagnetics, and atmospheric physics. It is the intent to review this handbook periodically to insure its completeness and currency. Users of this document are encouraged to report any errors discovered and any recommendations for changes or inclusions to: Commander, 1842 EEG/EEITE, Scott AFB IL 62225-6348.

5. Copies of Federal and Military Standards, Specifications and associated documents (including this handbook) listed in the Department of Defense Index of Specifications and Standards (DODISS) should be obtained from the DOD Single Stock Point: Commanding Officer, Naval Publications and Forms Center, 5801 Tabor Avenue, Philadelphia PA 19120. Single copies may be obtained on an emergency basis by calling (ALTOVON) 442-3321 or Area Code (215)-697-3321. Copies of industry association documents should be obtained from the sponsor. Copies of all other listed documents should be obtained from the contracting activity or as directed by the contracting officer.

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PREFACE

This volume is one of a two-volume series which sets forth the grounding, bonding, and shielding theory for communications electronics (C-E) equipments and facilities. Grounding, bonding, and shielding are complex subjects about which in the past there has existed a good deal of misunderstanding. The subjects themselves are interrelated and involve considerations of a wide range of topics from electrochemistry and metallurgy to electromagnetic field theory and atmospheric physics. These two volumes reduce these varied considerations into a usable set of principles and practices which can be used by all concerned with, and responsible for, the safety and effective operation of complex C-E systems. Where possible, the principles are reduced to specific steps. Because of the large number of interrelated factors, specific steps cannot be set forth for every possible situation. However, once the requirements and constraints of a given situation are defined, the appropriate steps for solution of the problem can be formulated utilizing the principles set forth.

Both volumes (Volume I, Basic Theory and Volume II, Applications) implement the Grounding, Bonding, and Shielding requirements of MIL-STD-188-124A which is mandatory for use within the Department of Defense. The purpose of this standard is to ensure the optimum performance of ground-based telecommunications equipment by reducing noise and providing adequate protection against power system faults and lightning strikes.

This handbook emphasizes the necessity for including considerations of grounding, bonding, and shielding in all phases of design, construction, operation, and maintenance of electronic equipment and facilities. Volume I, Basic Theory, develops the principles of personnel protection, fault protection, lightning protection, interference reduction, and EMP protection for C-E facilities. In addition, the basic theories of earth connections, signal grounding, electromagnetic shielding, and electrical bonding are presented. The subjects are not covered independently, rather they are considered from the standpoint of how they influence the design of the earth electrode subsystem of a facility, the selection of ground reference networks for equipments and structures, shielding requirements, facility and equipment bonding practices, etc. Volume I also provides the basic background of theory and principles that explain the technical basis for the recommended practices and procedures; illustrates the necessity for care and thoroughness in implementation of grounding, bonding, and shielding; and provides supplemental information to assist in the solution of those problems and situations not specifically addressed.

In Volume II, Applications, the principles and theories, including RED/BLACK protection, are reduced to the practical steps and procedures which are to be followed in structural and facility development, electronic engineering, and in equipment development. These applications should assure personnel equipment and structural safety, minimize electromagnetic interference (EMI) problems in the final operating system; and minimize susceptibility to and generation of undesirable emanations. The emphasis in Volume II goes beyond development to assembly and construction, to installation and checkout, and to maintenance for long term use.

Four appendices are provided as common elements in both volumes. Appendix A is a glossary of selected words and terms as they are used herein. If not defined in the glossary, usage is in accordance with Federal Standard 1037, Glossary of Telecommunication Terms. Appendix B is a supplemental bibliography containing selected references intended to supply the user with additional material. Appendix C contains the table of contents for the other volume. Appendix D contains the index for the two-volume set.

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## METRIC CONVERSION FACTORS

## Approximate Conversions to Metric Measures

Symbol	What You Know	Multiply by	To find	Symbol
<b>LENGTH</b>				
in	inches	2.5	centimeters	cm
ft	feet	30	centimeters	cm
y	yards	0.9	meters	m
m	miles	1.6	kilometers	km
<b>AREA</b>				
sq in	square inches	6.5	square centimeters	cm <sup>2</sup>
sq ft	square feet	0.09	square meters	m <sup>2</sup>
sq yd	square yards	0.8	square meters	m <sup>2</sup>
sq mi	square miles	2.6	square kilometers	km <sup>2</sup>
ac	acres	0.4	hectares	ha
<b>MASS (weight)</b>				
oz	ounces	28	grams	g
lb	pounds	0.45	kilograms	kg
sh ton	short tons (2000 lb)	0.9	tonnes	t
<b>VOLUME</b>				
imp gal	imperial gallons	4	liters	l
US gal	US gallons	3.8	liters	l
cu ft	cubic feet	0.03	cubic meters	m <sup>3</sup>
cu yd	cubic yards	0.76	cubic meters	m <sup>3</sup>
<b>TEMPERATURE (exact)</b>				
F	Fahrenheit temperature	5/9 (after subtracting 32)	Celsius temperature	°C

## Approximate Conversions from Metric Measures

Symbol	What You Know	Multiply by	To find	Symbol
<b>LENGTH</b>				
mm	millimeters	0.04	inches	in
cm	centimeters	0.4	inches	in
m	meters	3.3	feet	ft
km	kilometers	0.6	miles	mi
ha	hectares	2.5	acres	ac
<b>MASS (weight)</b>				
g	grams	0.005	ounces	oz
kg	kilograms	2.2	pounds	lb
t	tonnes (1000 kg)	1.1	short tons	sh ton
<b>VOLUME</b>				
l	liters	0.03	cubic feet	cu ft
ml	milliliters	0.03	cubic inches	cu in
l	liters	1.06	quarts	qt
l	liters	0.26	gallons	gal
m <sup>3</sup>	cubic meters	35	cubic feet	cu ft
m <sup>3</sup>	cubic meters	1.3	cubic yards	cu yd
<b>TEMPERATURE (exact)</b>				
°C	Celsius temperature	9/5 (then add 32)	Fahrenheit temperature	°F





## CHAPTER 1

## FACILITY GROUND SYSTEM

1.1 GENERAL.

1.1.1 This handbook addresses the practical considerations for engineering of grounding systems, subsystems, and other components of ground networks. Electrical noise reduction is discussed as it relates to the proper installation of ground systems. Power distribution systems are covered to the degree necessary to understand the interrelationships between grounding, power distribution, and electrical noise reduction.

1.1.2 The information provided in this handbook primarily concerns grounding, bonding, and shielding of fixed plant telecommunications-electronics facilities; however, it also provides basic guidance in the grounding of deployed transportable communications/electronics equipment.

1.1.3 Grounding, bonding, and shielding are approached from a total system concept, which comprises four basic subsystems in accordance with current Department of Defense (DOD) guidance. These subsystems are as follows:

- a. An earth electrode subsystem.
- b. A lightning protection subsystem.
- c. A fault protection subsystem.
- d. A signal reference subsystem.

1.2 APPLICATION. This handbook provides technical information for the engineering and installation of military communications systems related to the background and practical aspects of installation practices applicable to grounding, bonding, and shielding. It also provides the latest concepts on communications systems grounding, bonding, and shielding installation practices as a reference for military communications installation personnel.

1.3 DEFINITIONS. A glossary of unique terms used in this handbook is provided in Appendix A. All other terms and definitions used in this handbook conform to those contained in Joint Chiefs of Staff Publication No. 1. (JCS Pub 1), FED-STD-1037, MIL-STD-463, and the Institute of Electrical and Electronics Engineers (IEEE) dictionary.

1.4 REFERENCED DOCUMENTS. Publications related to the subject material covered in the text of this handbook are listed in Appendix B. The list includes publications referenced in the text and those documents that generally pertain to subjects contained in the handbook but are not necessarily addressed specifically.

1. DESCRIPTION. The ground system serves three primary functions which are listed below. A good ground system must receive periodic inspection and maintenance to retain its effectiveness. Continued or periodic maintenance is aided through adequate design, choice of materials, and proper installation techniques to ensure that ground subsystems resist deterioration or inadvertent destruction and thus require minimal repair to retain their effectiveness throughout the life of the facility.

a. Personnel safety. Personnel safety is provided by low-impedance grounding and bonding between equipment, metallic objects, piping, and other conductive objects, so that currents due to faults or lightning do not result in voltages sufficient to cause a shock hazard.

b. Equipment and facility protection. Equipment and facility protection is provided by low-impedance grounding and bonding between electrical services, protective devices, equipment, and other conductive objects, so that faults or lightning currents do not result in hazardous voltages within the facility. Also, the proper operation of overcurrent protective devices is frequently dependent upon low-impedance fault current paths.

c. Electrical noise reduction. Electrical noise reduction is accomplished on communication circuits by ensuring that (1) minimum voltage potentials exist between communications-electronics equipments, (2) the impedance between signal ground points throughout the facility to earth is minimal, and (3) that interference from noise sources is minimized.

1.5.1 Facility Ground System. All telecommunications and electronic facilities are inherently related to earth by capacitive coupling, accidental contact, and intentional connection. Therefore, ground must be looked at from a total system viewpoint, with various subsystems comprising the total facility ground system. The facility ground system forms a direct path of known low impedance between earth and the various power, communications, and other equipments that effectively extends in approximation of ground reference throughout the facility. The facility ground system is composed of an earth electrode subsystem, lightning protection subsystem, fault protection subsystem, and signal reference subsystem.

a. Earth electrode subsystem. The earth electrode subsystem consists of a network of earth electrode rods, plates, mats, or grids and their interconnecting conductors. The extensions into the building are used as the principal ground point for connection to equipment ground subsystems serving the facility. Ground reference is established by electrodes in the earth at the site or installation. The earth electrode subsystem includes the following: (1) a system of buried, driven rods interconnected with bare wire that normally form a ring around the building; or (2) metallic pipe systems, i.e., water, gas, fuel, etc., that have no insulation joints; or (3) a ground plane of horizontal buried wires. Metallic pipe systems shall not be used as the sole earth electrode subsystem. Resistance to ground should be obtained from the appropriate authority if available or determined by testing. For EMP considerations, see Chapter 10.

b. Lightning protection subsystem. The lightning protection subsystem provides a nondestructive path to ground for lightning energy contacting or induced in facility structures. To effectively protect a building, mast, tower, or similar self-supporting objects from lightning damage, an air terminal (lightning rod) of adequate mechanical strength and electrical conductivity to withstand the stroke impingement must be provided. An air terminal will intercept the discharge to keep it from penetrating the nonconductive outer coverings of the structure, and prevent it from passing through devices likely to be damaged or destroyed. A

low-impedance path from the air terminal to earth must also be provided. These requirements are met by either (1) an integral system of air terminals, roof conductors, and down conductors securely interconnected to provide the shortest practicable path to earth; or (2) a separately mounted shielding system, such as a metal mast or wires (which act as air terminals) and down conductors to the earth electrode subsystem.

c. **Fault protection subsystem.** The fault protection subsystem ensures that personnel are protected from shock hazard and equipment is protected from damage or destruction resulting from faults that may develop in the electrical system. It includes deliberately engineered grounding conductors (green wires) which are provided throughout the power distribution system to afford electrical paths of sufficient capacity, so that protective devices such as fuses and circuit breakers installed in the phase or hot leads can operate promptly. If at all possible the equipment fault protection conductors should be physically separate from signal reference grounds except at the earth electrode subsystem. The equipment fault protection subsystem provides grounding of conduits for signal conductors and all other structural metallic elements as well as the cabinets or racks of equipment.

d. **Signal reference subsystem.** The signal reference subsystem establishes a common reference for C-E equipments, thereby also minimizing voltage differences between equipments. This in turn reduces the current flow between equipments and also minimizes or eliminates noise voltages on signal paths or circuits. Within a piece of equipment, the signal reference subsystem may be a bus bar or conductor that serves as a reference for some or all of the signal circuits in the equipment. Between equipments, the signal reference subsystem will be a network consisting of a number of interconnected conductors. Whether serving a collection of circuits within an equipment or serving several equipments within a facility, the signal reference network will in the vast majority of cases be a multiple point/equipotential plane but could also, in some cases, be a single point depending on the equipment design, the facility, and the frequencies involved.

**1.5.2 Grounding and Power Distribution Systems.** For safety reasons, both the MIL-STD-188-124A and the National Electrical Code (NEC) require the electrical power systems and equipments be intentionally grounded; therefore, the facility ground system is directly affected by the proper installation and maintenance of the power distribution systems. The intentional grounding of electrical power systems minimizes the magnitude and duration of overvoltages on an electrical circuit, thereby reducing the probability of personnel injury, insulation failure, or fire and consequent system, equipment, or building damage.

a. Alternating currents in the facility ground system are primarily caused as a result of improper ac wiring, simple mistakes in the ac power distribution system installation, or as a result of power faults. To provide the desired safety to personnel and reduce equipment damage, all 3-phase wye wiring to either fixed or transportable communication facilities shall be accomplished by the 5-wire or conductor distribution system consisting of three phase or "hot" leads, one neutral lead and one grounding (green) conductor. A single building receiving power from a single source requires the ac neutral be grounded to the earth electrode subsystem on the source side of the first service disconnect or service entrance panel as well to a ground terminal at the power source (transformer, generator, etc.). This neutral shall not be grounded at any point within the building or on the load side of the service entrance panel. The grounding of all C-E equipment within the building is accomplished via the grounding (green) conductor which is bonded to the neutral bus in the source side of the service entrance panel and, in turn, grounded to the earth electrode subsystem. In addition to the three phase or "hot" leads and the neutral (grounded) conductor, a fifth wire is employed to interconnect the facility earth electrode subsystem with the ground terminal at the power source.

To eliminate or reduce undesired noise or hum, multiple facilities supplied from a single source shall ground the neutral only at the power source and not to the earth electrode subsystem at the service entrance point. Care should be taken to ensure the neutral is not grounded on the load side of the first disconnect service or at any point within the building. The grounding (green) conductor in this case is not bonded to the neutral bus in the service disconnect panel. It is, however, bonded to the facility earth electrode subsystem at the service entrance panel. The fifth wire shall be employed to interconnect the earth electrode subsystem with the ground terminal at the power source.

The secondary power distribution wiring for a 240 volt single phase system consists of two phase or "hot" leads, a neutral (grounded) and a grounding (green) conductor while the three conductor secondary power distribution system is comprised of one phase, one neutral, and one grounding lead. In both cases, the neutral shall not be grounded on the load side of the first service disconnect. It shall, however, be grounded to the ground terminal at the power source and to the earth electrode subsystem if one power source supplies power only to a single building.

The ac wiring sequence (phase, neutral, and equipment fault protection) must be correct all the way from the main incoming ac power source to the last ac load, with no reversals between leads and no interconnection between neutral and ground leads. Multiple ac neutral grounds and reversals between the ac neutral and the fault protection subsystem will generally result in ac currents in all ground conductors to varying degrees. The NEC recognizes and allows the removal or relocation of grounds on the green wire which cause circulating currents. (Paragraph 250-21(b) of the NEC refers.) Alternating current line filters also cause some ac currents in the ground system when distributed in various areas of the facility; this is due to some ac current passing through capacitors in the ac line filters when the lines are filtered to ground. Power line filters should not induce more than 30 milliamperes of current to the fault protection subsystem.

b. Dc power equipment has been found to be a significant electrical noise source that can be minimized through proper configuration of the facility, the physical and electrical isolation of the dc power equipment from communications equipment, and filtering of the output. Certain communications equipment with inverter or switching type power supplies also cause electrical noise on the dc supply leads and the ac input power leads. This noise can be minimized by the use of decentralizing filters at or in the equipment. The location, number, and termination of the dc reference ground leads are also important elements in providing adequate protection for dc systems and, at the same time, minimizing electrical noise and dc currents in the ground system.

1.5.3 Electrical Noise in Communications Systems. Interference-causing signals are associated with time-varying, repetitive electromagnetic fields and are directly related to rates of change of currents with time. A current-changing source generates either periodic signals, impulse signals, or a signal that varies randomly with time. To cause interference, a potentially interfering signal must be transferred from the point of generation to the location of the susceptible device. The transfer of noise may occur over one or several paths. There are several modes of signal transfer (i.e., radiation, conduction, and inductive and capacitive coupling).

## 1.6 BONDING, SHIELDING, AND GROUNDING RELATIONSHIP.

a. The simple grounding of elements of a communications facility is only one of several measures necessary to achieve a desired level of protection and electrical noise suppression. To provide a low-impedance path for (1) the flow of ac electrical current to/from the equipment and (2) the achievement of an effective grounding system, various conductors, electrodes, equipment, and other metallic objects must be joined or bonded together. Each of these bonds should be made so that the mechanical and electrical properties of the path are determined by the connected members and not by the interconnection junction. Further, the joint must maintain its properties over an extended period of time, to prevent progressive degradation of the degree of performance initially established by the interconnection. Bonding is concerned with those techniques and procedures necessary to achieve a mechanically strong, low-impedance interconnection between metal objects and to prevent the path thus established from subsequent deterioration through corrosion or mechanical looseness.

b. The ability of an electrical shield to drain off induced electrical charges and to carry sufficient out-of-phase current to cancel the effects of an interfering field is dependent upon the shielding material and the manner in which it is installed. Shielding of sensitive electrical circuits is an essential protective measure to obtain reliable operation in a cluttered electromagnetic environment. Solid, mesh, foil, or stranded coverings of lead, aluminum, copper, iron, and other metals are used in communications facilities, equipment, and conductors to obtain shielding. These shields are not fully effective unless proper bonding and grounding techniques are employed during installation. Shielding effectiveness of an equipment or subassembly enclosure depends upon such considerations as the frequency of the interfering signal, the characteristics of the shielding material, and the number and shapes of irregularities (openings) in the shield.

## 1.7 GROUNDING SAFETY PRACTICES.

a. It is essential that all personnel working with Communications-Electronics (C-E) equipment and supporting systems and facilities strictly observe the rules, procedures, and precautions applicable to the safe installation, operation, and repair of equipment and facilities. All personnel must be constantly alert to the potential hazards and dangers presented and take all measures possible to reduce or eliminate accidents.

b. Safety precautions in the form of precisely worded and illustrated danger or warning signs shall be prominently posted in conspicuous places, to prevent personnel from making accidental contact with high-voltage sources such as power lines, antennas, power supplies, or other places where uninsulated contacts present the danger of electrical shock or short circuits. Signs shall also warn of the dangers of all forms of radiation hazards, acids, and chemical inhalation, plus all other potential sources of personnel danger. Power cutoff features built into the equipment must be used in strict adherence to the intended use.

c. During the installation of equipment, warning tags are used to note the existence of potential danger when individual circuits or stages are being checked out. The tags should contain appropriate information to alert all personnel of the dangers involved and specific restrictions as to the use of the equipment. The equipment being installed shall be appropriately tagged in accordance with the directives of the local safety officer, equipment manufacturer, or other responsible agent.

d. Installation personnel, when working with equipment having high-voltage devices, must ensure that the devices are grounded and that the high-voltage circuits have been disconnected or turned off. Do not rely solely on the presence of interlock switches for protection from electrical shock.

## CHAPTER 2

## EARTH ELECTRODE SUBSYSTEM

2.1 OBJECTIVES.

Earth grounding is defined as the process by which an electrical connection is made to the earth. The earth electrode subsystem is that network of interconnected rods, wires, pipes, or other configuration of metals which establishes electrical contact between the elements of the facility and the earth. This system should achieve the following objectives:

- a. Provide a path to earth for the discharge of lightning strokes in a manner that protects the structure, its occupants, and the equipment inside.
- b. Restrict the step-and-touch potential gradient in areas accessible to persons to a level below the hazardous threshold even under lightning discharge or power fault conditions.
- c. Assist in the control of noise in signal and control circuits by minimizing voltage differentials between the signal reference subsystems of separate facilities.

2.1.1 Lightning Discharge. A lightning flash is characterized by one or more strokes with typical peak current amplitudes of 20 kA or higher. In the immediate vicinity of the point of entrance of the stroke current into the earth, hazardous voltage gradients can exist along the earth's surface. Ample evidence (2-1)\* exists to show that such gradients are more than adequate to cause death. It is thus of great importance that the earth electrode subsystem be configured in a manner that minimizes these gradients. The lower the resistance of the earth connection, the lower the peak voltage and consequently the less severe the surface gradients. Even with low resistance earth electrode systems, the current paths should be distributed in a way that minimizes the gradients over the area where personnel might be present.

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\* Referenced documents are listed in the last section of each chapter.

**2.1.2 Fault Protection.** In the event of transformer failure (e.g., disconnect between neutral and ground or line to ground faults) or any failure between the service conductor(s) and grounded objects in the facility, the earth electrode subsystem becomes a part of the return path for the fault current. A low resistance assists in fault clearance; however, it does not guarantee complete personnel protection against hazardous voltage gradients which are developed in the soil during high current faults. Adequate protection generally requires the use of ground grids or meshes designed to distribute the flow of current over an area large enough to reduce the voltage gradients to safe levels. The neutral conductor at the distribution transformer must therefore be connected to the earth electrode subsystem to ensure that a low resistance is attained for the return path. (Paragraph 5.1.1.2.5.1 of MIL-STD-188-124A refers.) Ground fault circuit interrupters on 120 volt single phase 15 and 20 ampere circuits will provide personnel protection against power faults and their use is therefore highly recommended.

**2.1.3 Noise Reduction.** The earth electrode subsystem is important for the minimization of electromagnetic noise (primarily lower frequency) within signal circuits caused as a result of stray power currents. For example, consider a system of two structures located such that separate earth electrode subsystems are needed as shown in Figure 2-1. If stray currents (such as may be caused by an improperly grounded ac system, dielectric leakage, high resistance faults, improperly returned dc, etc.) are flowing into the earth at either location, then a voltage differential will likely exist between the grounding networks within each facility.

Currents originating from sources outside the structures can also be the cause of these noise voltages. For example, high voltage substations are frequent sources of large power currents in the earth. Such currents arise from leakage across insulators, through cable insulation, and through the stray capacitance which exists between power lines and the earth. These currents flowing through the earth between the two sites will generate a voltage difference between the earth connections of the two sites in the manner illustrated by Figure 2-2.

Any interconnecting wires or cables will have these voltages applied across the span which will cause currents to flow in cable shields and other conductors. As shown in Chapter 6, such intersite currents can induce common-mode noise voltages into interconnected earth electrode subsystems.

**2.1.4 Summary of Requirements.** Table 2-1 summarizes the purpose, requirements, and resulting design factors for earth connections of the lightning protection subsystem, the fault protection subsystem, the signal reference subsystem, and the ac distribution system neutral (grounded) conductor and safety ground (grounding) conductor. Refer to Article 100 - Definitions of the NEC for additional information on grounding and grounded conductors [2-2].



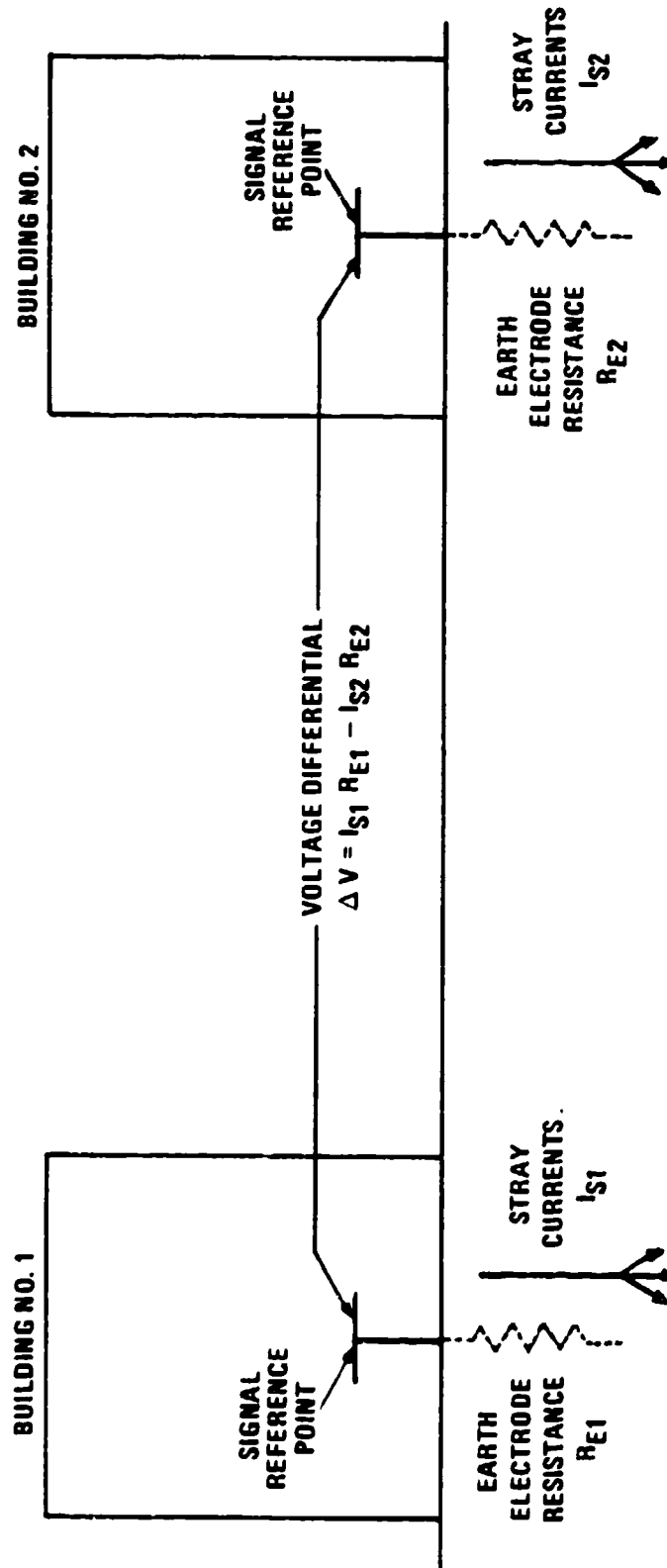


Figure 2-1. Voltage Differentials Arising From Unequal Earth Electrode Resistances and Unequal Stray Currents

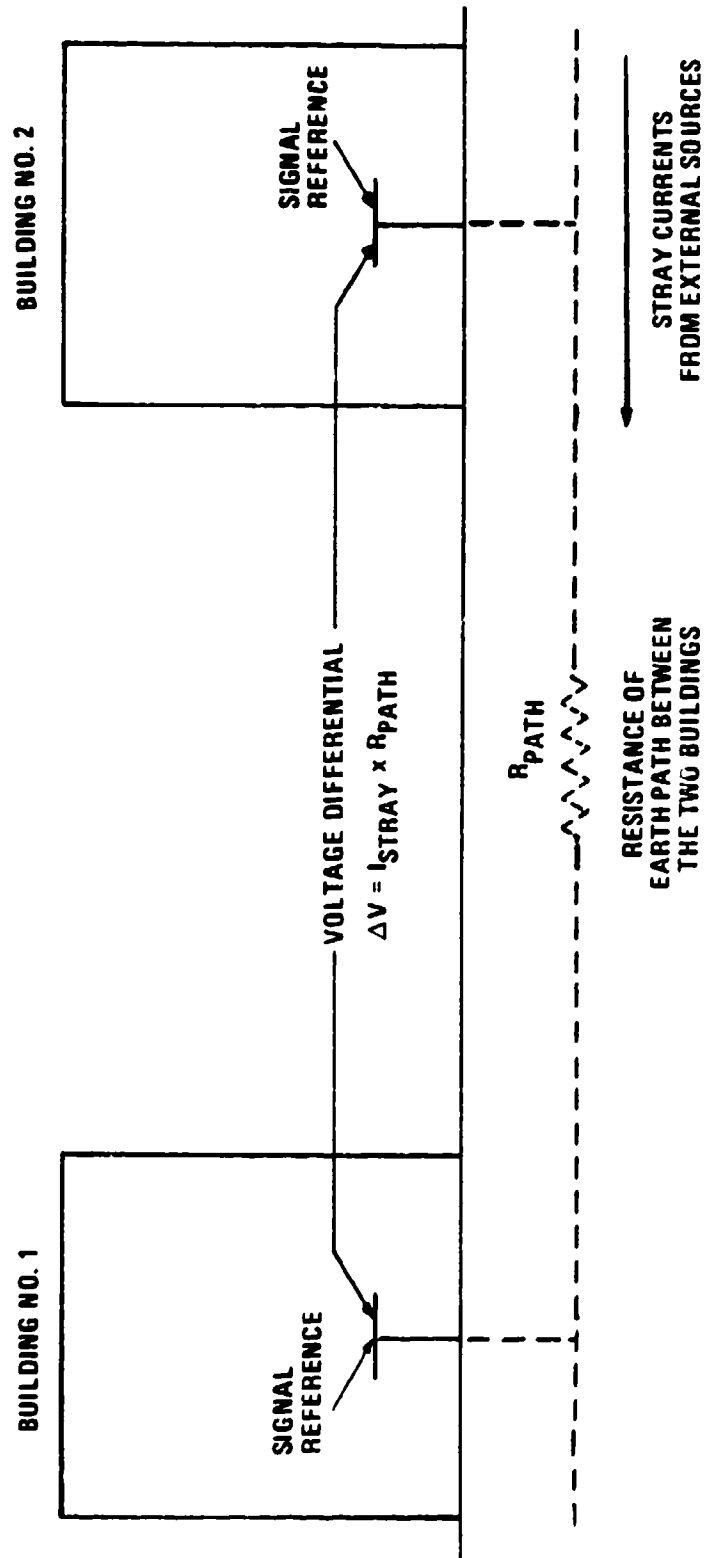


Figure 2-2. Voltage Differentials Between Structures Resulting From Stray Ground Currents

## 2.2 RESISTANCE REQUIREMENTS.

2.2.1 General. The basic measure of effectiveness of an earth electrode is the value in ohms of the resistance to earth at its input connection. Because of the distributed nature of the earth volume into which electrical energy flows, the resistance to earth is defined as the resistance between the point of connection and a very distant point on the earth (see Section 2.4). Ideally, the earth electrode subsystem provides a zero resistance between the earth and the point of connection. Any physically realizable configuration, however, will exhibit a finite resistance to earth. The economics of the design of the earth electrode subsystem involve a trade-off between the expense necessary to achieve a very low resistance and the satisfaction of minimum system requirements. This subsystem shall also interconnect all driven electrodes and underground metal objects of the facilities including the emergency power plant. Underground metallic pipes entering the facility shall also be bonded to the earth electrode subsystem.

2.2.2 Resistance to Earth. Metal underground water pipes typically exhibit a resistance to earth of less than three ohms. Other metal elements in contact with the soil such as the metal frame of the building, underground gas piping systems, well casings, other piping and/or buried tanks, and concrete-encased steel reinforcing bars or rods in underground footings or foundations generally exhibit a resistance substantially lower than 25 ohms.

2.2.2.1 National Electrical Code Requirements. For the fault protection subsystem, the NEC (2-2) states in Article 250 that a single electrode consisting of a rod, pipe or plate which does not have a resistance to ground of 25 ohms or less shall be augmented by one additional made electrode. Although the language of the NEC clearly implies that electrodes with resistances as high as 25 ohms are to be used only as a last resort, this 25 ohm limit has tended to set the norm for grounding resistance regardless of the specific system needs. The 25 ohm limit is reasonable or adequate for application to private homes and other lower powered type facilities.

2.2.2.2 Department of Defense Communications Electronics Requirements. The above criteria however, is not acceptable for C-E facilities when consideration is given to the large investments in personnel and equipment. A compromise of cost versus protection against lightning, power faults, or EMP has led to establishment of a design goal of 10 ohms for the earth electrode subsystem (EES) in MIL-STD-188-124A. The EES designed in MIL-STD-188-124A specifies a ring ground around the periphery of the facility to be protected. With proper design and installation of the EES, the design goal of 10 ohms should be attained at reasonable cost. At locations where the 10 ohms has not been attained due to high soil resistivity, rock formations, or other terrain features, alternate methods listed in Paragraph 2.9 shall be considered for reducing the resistance to earth.

2.2.3 Lightning Requirements. For lightning protection, it also is difficult to establish a definite grounding resistance necessary to protect personnel. The current which flows in a direct lightning stroke may vary from several hundred amperes to as much as 300 thousand amperes. Such currents through even one ohm of resistance can theoretically produce hazardous potentials. It is impractical to attempt to reduce the resistance of a facility to earth to a value low enough to absolutely prevent the development of these potentials. Techniques other than simply achieving an extremely low resistance to ground must therefore be employed to protect personnel and equipment inside a structure from the hazards produced by a direct stroke. Experience has shown that a grounding resistance of ten ohms gives fairly reliable lightning protection to buildings, transformers, transmission lines, towers, and other exposed structures. At some sites, resistances as low as one ohm or less can be achieved economically. The lower the resistance, the greater the protection; therefore, attempts should be made to reduce the resistance to the lowest practical value.

Table 2-1  
Facility Ground System: Purposes, Requirements, and Design Factors

Subsystem	Purpose	Requirements	Design Factors
Lightning Protection	Dissipate lightning energy in earth.	Multiple connections to earth electrode subsystem, high peak power transfer capability, low impulse impedance to minimize magnitude of transient potentials.	Lightning protection subsystem must be sized to dissipate energy in a lightning pulse (worst case) without producing hazardous voltages or damage to itself.
Fault Protection	Provide fault current path to operate equipment breakers, blow fuses, etc.	Low resistance in the return path for fault current, maintain voltage of equipment enclosures near earth potential.	Resistance should be low enough to permit operation of facility over-current devices when faults occur.
Signal Reference	Reduce noise in signal circuits, provide leakage path for static charges, establish voltage reference.	Establish reference potential for signal voltages, provide sink for static charge.	Fault currents and lightning protection system currents normally should not flow in the signal reference network; earth connection should not degrade signal quality.
Earth Electrode	Low resistance path to earth.	Provides link for lightning protection, fault protection and signal reference subsystems to earth.	Installed around periphery of building or tower to be protected.

## 2.3 SOIL RESISTIVITY.

2.3.1 General. The resistivities of the soil and rock in which the earth electrode subsystem is buried, constitute the basic constraint on the achievement of a low resistance contact with earth. The resistance of an earth electrode subsystem can in general be calculated with formulas which are based upon the general resistance formula.

$$R = \rho \frac{l}{A} = \frac{E}{I} \quad (2-1)$$

where  $\rho$  is the resistivity of the conducting material,  $l$  is the length of the path for current flow in the earth,  $A$  is the cross-sectional area of the conducting path,  $I$  is the current into the electrode, and  $E$  is the voltage of the electrode measured with respect to infinity. It will be shown later in this chapter that if the soil resistivity is known, the resistance of the connection provided by the more common electrode configurations can be readily determined.

The soils of the earth consist of solid particles and dissolved salts. Electrical current flows through the earth primarily as ion movement; the ionic conduction is heavily influenced by the concentration and kinds of salts in the moisture in the soil. Ionic disassociation occurs when salts are dissolved, and it is the movement of these ions under the influence of electrical potential which enable the medium to conduct electricity.

Resistivity is defined in terms of the electrical resistance of a cube of homogeneous material. The resistance of a homogeneous cube, as measured across opposite faces, is proportional to the resistivity and inversely proportional to the length of one side of the cube. The resistance is

$$R = \rho \frac{L}{A} = \rho \frac{L}{L^2} = \left( \frac{\rho}{L} \right) \text{ ohms} \quad (2-2)$$

where  $\rho$  = resistivity of the material, ohms - (unit-of-length);

$L$  = length of one side of the cube, (unit-of-length), and

$A$  = area of one face of the cube, (unit-of-length)<sup>2</sup>.

Common units of resistivity are ohm-cm and ohm-m.

2.3.2 Typical Resistivity Ranges. A broad variation of resistivity occurs as a function of soil types, and classification of the types of soils at a potential site for earth electrodes is needed by the designer. Table 2-2 permits a quick estimate of soil resistivity, while Table 2-3 lists measured resistivity values from a variety of sources. Tables 2-2 and 2-3 indicate that ranges of one or two orders of magnitude in values of resistivity for a given soil type are to be expected.

2.3.3 Environmental Effects. In addition to the variation with soil types, the resistivity of a given type of soil will vary several orders of magnitude with small changes in the moisture content, salt concentration, and soil temperature. It is largely these variations in soil environment that cause the wide range of values for each soil type noted in Tables 2-2 and 2-3. Figure 2-3 shows the variations observed in a particular soil as moisture, salt, and temperature were changed. The curves are intended only to indicate trends -- another type of soil would be expected to yield curves with similar shapes but different values.

The discontinuity in the temperature curve (Figure 2-3(b)), indicates that at below freezing temperatures the soil resistivity increased markedly. This undesirable temperature effect can be minimized by burying earth electrode subsystems below the frost line.

## 2.4 MEASUREMENT OF SOIL RESISTIVITY.

2.4.1 General. It is not always possible to ascertain with a high degree of certainty the exact type of soil present at a given site. Soil is typically rather nonhomogeneous; many types will be encountered at most locations. Even with the aid of borings and test samples and the use of Table 2-3, the resistivity estimate can easily be off by two or three orders of magnitude. When temperature and moisture variations are added to the soil type variations, it is evident that estimates based on Table 2-3 are not sufficiently accurate for design purposes. The only way to accurately determine the resistivity of the soil at a specific location is to measure it.

2.4.2 Measurement Techniques. The most commonly used field methods for determining soil resistivity employ the technique of injecting a known current into a given volume of soil, measuring the voltage drop produced by the current passing through the soil, and then determining the resistivity from a modified form of Equation 2-1.

2.4.2.1 One-Electrode Method. To illustrate the principles of this technique, first visualize a metal hemisphere buried in the earth as shown in Figure 2-4. In uniform earth, injected current flows radially from this hemispherical electrode. Equipotential surfaces are established concentric with the electrode and perpendicular to the radial directions of current flow. (Regardless of the shape of an electrode, it can be approximated as a hemispherical electrode if viewed from far enough away.) As the current flows from the hemisphere, the current density decreases with distance from the electrode because the areas of successive shells become larger and larger. The current density within the earth, at a given distance  $x$  from the center of the electrode is

$$i_x = \frac{I}{2\pi x^2} \text{ amperes per unit area,} \quad (2-3)$$

where

$I$  = current entering the electrode and

$2\pi x^2$  = area of the hemispherical shell with radius  $x$ .

At the point  $x$  the electric field strength can be obtained from Ohm's law:

$$\begin{aligned} e_x &= \rho i_x, \\ &= \frac{\rho I}{2\pi x^2} \text{ volts per unit length.} \end{aligned} \quad (2-4)$$

where  $\rho$  is resistivity of material.

Table 2-2

## Approximate Soil Resistivity (2-3)

Type of Soil	Resistivity		
	(ohm-m)	(ohm-cm)	(ohm-ft)
Wet Organic Soil	10	$10^3$	33
Moist Soil	$10^2$	$10^4$	330
Dry Soil	$10^3$	$10^5$	3300
Bed Rock	$10^4$	$10^6$	33000

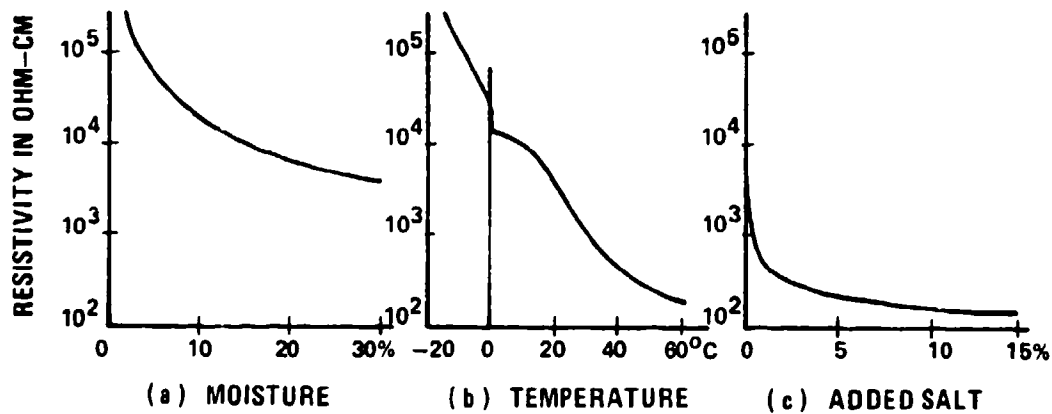


Figure 2-3. Typical Variations in Soil Resistivity as a Function of Moisture, Temperature and Salt Content (2-4)

Table 2-3

## Resistivity Values of Earthing Medium (2-5), (2-6), (2-7)

Medium	Resistivity		
	Minimum (ohm-cm)	Average (ohm-cm)	Maximum (ohm-cm)
Surface soils, loam, etc.	$10^2$		$5 \times 10^3$
Clay	$2 \times 10^2$		$10^4$
Sand and gravel	$5 \times 10^3$		$10^5$
Surface limestone	$10^4$		$10^6$
Limestones	$5 \times 10^2$		$4 \times 10^5$
Shales	$5 \times 10^2$		$10^4$
Sandstone	$2 \times 10^3$		$2 \times 10^5$
Granites, basalts, etc.		$10^6$	
Decomposed gneisses	$5 \times 10^3$		$5 \times 10^4$
Slates, etc.	$10^3$		$10^4$
Fresh Water Lakes		$2 \times 10^4$	$2 \times 10^7$
Tap Water	$10^3$		$5 \times 10^3$
Sea Water	20	$10^2$	$2 \times 10^2$
Pastoral, low hills, rich soil, typical of Dallas, Texas; Lincoln, Nebraska areas		$3 \times 10^3$	
Flat country, marshy, densely wooded typical of Louisiana near Mississippi River	$2 \times 10^2$	$10^4$	
Pastoral, medium hills and forestation, typical of Maryland, Pennsylvania, New York, exclusive of mountainous territory and seacoasts		$2 \times 10^4$	
Rocky soil, steep hills, typical of New England	$10^3$	$5 \times 10^4$	$10^5$
Sandy, dry, flat, typical of coastal country	$3 \times 10^4$	$5 \times 10^4$	$5 \times 10^5$
City, industrial areas		$10^5$	$10^6$



Table 2-3 (Continued)

Resistivity Values of Earthing Medium (2-5), (2-6), (2-7)

Medium	Resistivity		
	Minimum (ohm-cm)	Average (ohm-cm)	Maximum (ohm-cm)
Fills, ashes, cinders, brine, waste	$6 \times 10^2$	$2.5 \times 10^3$	$7 \times 10^3$
Clay, shale, gumbo, loam	$3 \times 10^2$	$4 \times 10^3$	$2 \times 10^4$
Same-with varying proportion of sand and gravel	$10^3$	$1.5 \times 10^4$	$10^5$
Gravel, sand stones with little clay or loam, granite	$5 \times 10^4$	$10^5$	$10^6$

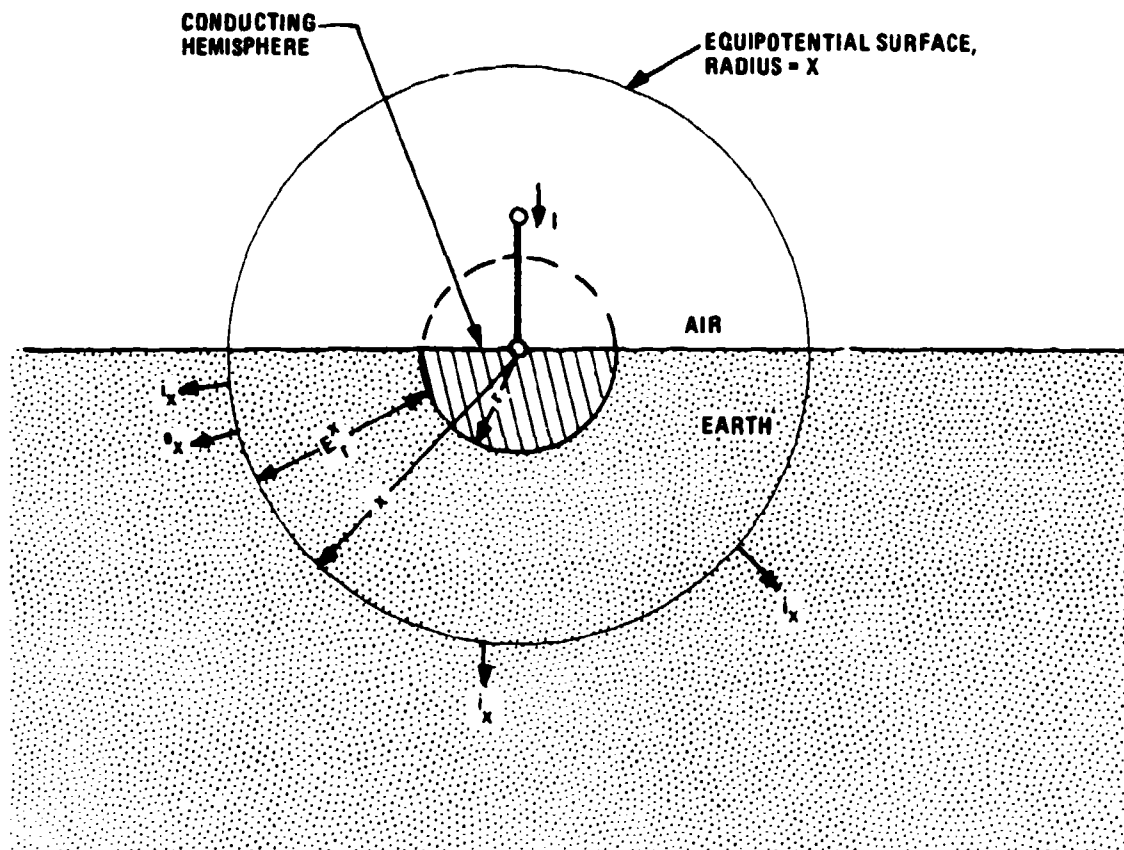


Figure 2-4. Current Flow From a Hemisphere in Uniform Earth

The voltage from the surface of the electrode to the point  $x$  is the line integral of  $e_x$  with the lower limit equal to the sphere's radius,  $r$ , and the upper limit equal to the distance,  $x$ :

$$E_r^x = \int_r^x e_x dx \quad (2-5)$$

$$= \frac{\rho I}{2\pi} \int_r^x \frac{dx}{x^2}$$

$$E_r^x = \frac{\rho I}{2\pi} \left[ \frac{1}{r} - \frac{1}{x} \right]$$

As  $x$  becomes very large,  $E$  is closely approximated as

$$E = \frac{\rho I}{2\pi r} \quad (2-6)$$

The resistance to the earth of the electrode is the resistance between the electrode and a very distant point; therefore

$$\begin{aligned} R &= \frac{E}{I} \\ &= \frac{\rho}{2\pi r} \end{aligned} \quad (2-7)$$

where:  $E$  = the voltage drop between the electrode and a point infinitely distant,  
 $I$  = the current entering the electrode,  
 $\rho$  = earth resistivity, and  
 $r$  = radius of hemisphere.

Rewriting Equation 2-7 as

$$\rho = 2\pi r R = 2\pi r \frac{E}{I} \quad (2-8)$$

shows that the resistivity can be determined by knowing  $r$ ,  $E$ , and  $I$ .

**2.4.2.2 Four-Terminal Method.** In the four-terminal method developed by the U.S. Bureau of Standards (2-8), four electrodes are inserted into the soil in a straight line with equal spacings. A known current is injected into the soil through the end electrodes and the voltage drop between the two inside electrodes is measured.

Consider four deeply buried spheres placed in a straight line, separated by a distance,  $a$ , as shown in Figure 2-5. Connection is made to the spheres by insulated conductors. Assume that a current,  $I$ , is introduced into one of the outermost spheres (No. 1) and flows out of the earth through the other (No. 4) outermost sphere. The voltage from the left hand (No. 2) to the right hand (No. 3) inner sphere can be viewed as resulting from a current flowing to infinity and another returning from infinity. The two resulting components of the voltage are (2-8)

$$V_1 = \frac{\rho I_i}{8\pi a}, \quad (2-9)$$

where

$I_i$  = input current,

and

$$V_2 = \frac{\rho I_o}{8\pi a}, \quad (2-10)$$

where

$I_o$  = output current.

But since

$$I_o = I_i,$$

the total potential  $V$  is

$$V = V_1 + V_2 \quad (2-11)$$

$$= \frac{\rho I}{4\pi a}.$$

If the probe depth,  $h$ , is less than the probe separation distance,  $a$ , the potential drop measured between the inner electrodes divided by the current measured into (or out of) one of the outer electrodes is (2-8):

$$R = \frac{V}{I} = \frac{\rho}{4\pi} \left( \frac{1}{a} + \frac{2}{\sqrt{a^2 + 4h^2}} - \frac{2}{\sqrt{4a^2 + 4h^2}} \right), \quad (2-12)$$

where:  $a$  = distance between four, equally spaced, in-line probes, and

$h$  = depth of burial of probes (insulated leads to surface).

If  $h \ll a$ , Equation 2-12 simplifies to

$$R = \frac{\rho}{2\pi a} \quad (2-13)$$

or

$$\rho = 2\pi a R. \quad (2-14)$$

Short rods provide an effective approximation to the buried sphere, particularly at distances large with respect to the depth of insertion.

The typical earth resistance test set contains a hand powered generator which can generate an ac signal at frequencies of 40 to 100 hertz or so. (Fifty or sixty hertz should not be used because errors may be produced by stray power currents in the soil. Direct current is not usually used because of polarization effects.) By adjusting the resistance of an internal double balanced bridge, the instrument provides a direct indication of the  $R$  required in Equation 2-14.

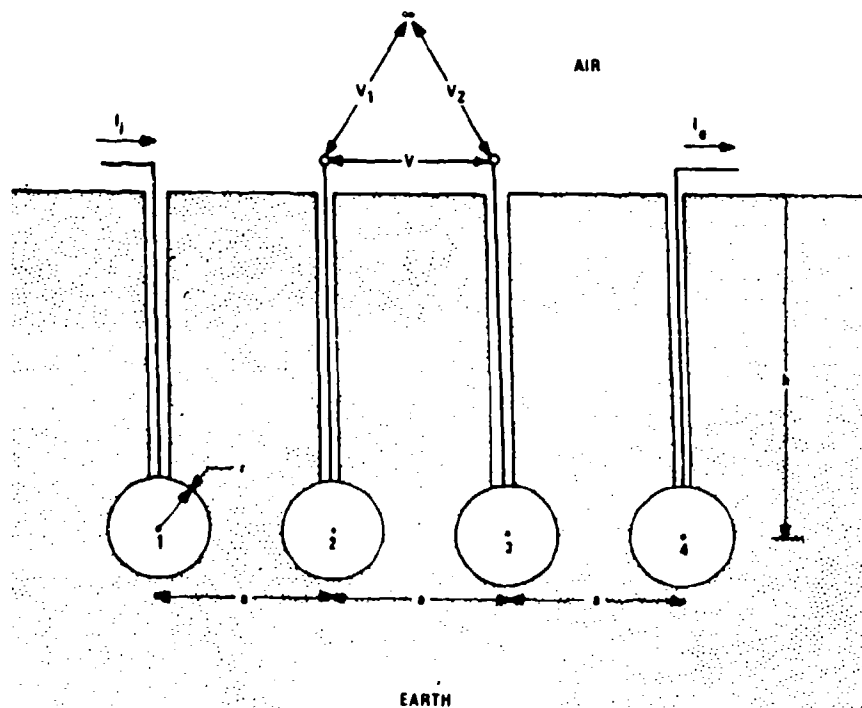


Figure 2-5. Idealized Method of Determining Soil Resistivity

## 2.5 TYPES OF EARTH ELECTRODE SUBSYSTEMS.

2.5.1 General. Earth electrode subsystems can be divided into two general types, the most preferable being a ring ground with 10-foot (3-meter) minimum length ground rods every 15 feet (4.5 meters). A second and less preferable type consists of a system of radials or grounds used when soil is rocky or has extremely high resistivity. At sites where soil resistivity varies from high to very high and frequent electrical storms are common, a combination of the two is recommended, i.e., a ring ground around the building (worst case--grid under building) extending 2 to 6 feet (0.6 to 1.8 meters) outside the drip line with radials or horizontal conductors extending to 125 feet (37.5 meters). With either system, resistance to earth and danger of arc over can be greatly reduced by bonding any large metal objects in the immediate area to the earth electrode subsystem. These include metal pipes, fuel tanks, grounded metal fences, and well casings.

2.5.2 Ground Rods. Vertically driven ground rods or pipes are the most common type of made electrode. Rods or pipes are generally used where bedrock is beyond a depth of 3 meters (10 feet). Ground rods are commercially manufactured in 1.27, 1.59, 1.90 and 2.54 cm (1/2, 5/8, 3/4 and 1 inch) diameters and in lengths from 1.5 to 12 meters (5 to 40 feet). For most applications, ground rods of 1.90 cm (3/4 inch) diameter, and length of 3.0 meters (10 feet), are used. Copper-clad steel ground rods are required because the steel core provides the strength to withstand the driving force and the copper provides corrosion protection and is compatible with copper or copper-clad interconnecting cables.

2.5.3 Buried Horizontal Conductors. Where bedrock is near the surface of the earth, the use of driven rods is impractical. In such cases, horizontal strips of metal, solid wires, or stranded cables buried 0.48 to 0.86 meters (18 to 36 inches) deep may be used effectively. With long strips, reactance increases as a factor of the length with a consequent increase in impedance. A low impedance is desirable for minimizing lightning surge voltages. Therefore, several wires, strips, or cables arranged in a star pattern, with the facility at the center, is preferable to one long length of conductor.

2.5.4 Grids. Grid systems, consisting of copper cables buried about 15.24 cm (6 inches) in the ground and forming a network of squares, are used to provide equipotential areas throughout the facility area. Such a system usually extends over the entire area. The spacing of the conductors, subject to variation according to requirements of the installation, may normally be 0.6 to 1.2 meters (2 to 4 feet) between cables. The cables must be bonded together at each crossover.

Grids are generally required only in antenna farms or substation yards and other areas where very high fault currents are likely to flow into the earth and hazardous step potentials may exist (see Section 2.8.1.2.3) or soil conditions prohibit installation of other ground systems. Antenna counterpoise systems shall be installed in accordance with guidance requirements of the manufacturer.

2.5.5 Plates. Rectangular or circular plate electrodes should present a minimum of 0.09 square meters (2 square feet) of surface contact with the soil. Iron or steel plates should be at least 0.64 cm (1/4 inch) thick and nonferrous metals should be at least 0.15 cm (0.06 inches) thick. A burial depth of 1.5 to 2.4 meters (5 to 8 feet) below grade should be maintained. This system is considered very expensive for the value produced and generally not recommended.

**2.5.6 Metal Frameworks of Buildings.** The metal frameworks of buildings may exhibit a resistance to earth of less than 10 ohms, depending upon the size of the building, the type of footing, and the type of subsoil at a particular location. Buildings that rest on steel pilings in particular may exhibit a very low resistance connection to earth. For this low resistance to be used advantageously, it is necessary that all elements of the framework be bonded together.

**2.5.7 Water Pipes.** Metal underground pipes have traditionally been relied upon for grounding electrodes. The resistance to earth provided by piping systems is usually quite low because of the extensive contact made with soil. Municipal water systems in particular establish contact with the soil over wide areas. For water pipes to be effective, any possible discontinuities must be bridged with bonding jumpers. The NEC requires that any water metering equipment and service unions be bypassed with a jumper not less than that required for the grounding connector.

However, stray or fault currents flowing through the piping network into the earth can present a hazard to workmen making repairs or modifications to the water system. For example, if the pipes supplying a building are disconnected from the utility system for any reason, that portion connected to the building can rise to a hazardous voltage level relative to the rest of the piping system and possibly with respect to the earth. In particular, if the resistance that is in contact with the soil near the building happens to be high, a break in the pipe at even some distance from the building may pose a hazardous condition to unsuspecting workmen. Some water utilities are inserting non-conductive couplings in the water mains at the point of entrance to buildings to prevent such possibilities. For these reasons, the water system should not be relied upon as a safe and dependable earth electrode for a facility and should be supplemented with at least one other ground system.

**2.5.8 Incidental Metals.** There may be a number of incidental, buried, metallic objects in the vicinity of the earth electrode subsystem. These objects should be connected to the system to reduce the danger of potential differences during lightning or power fault conditions; their connection will also reduce the resistance to earth of the earth electrode subsystem. Such additions to the earth electrode subsystem should include the rebar in concrete footings, buried tanks, and piping.

**2.5.9 Well Casings.** Well casing can offer a low resistance contact with the earth. In some areas, steel pipe used for casing in wells can be used as a ground electrode. Where wells are located on or near a site, the resistance to earth of the casing should be measured and, if below 10 ohms, the well casing can be considered for use as a ground electrode.

## 2.6 RESISTANCE PROPERTIES.

### 2.6.1 Simple Isolated Electrodes.

2.6.1.1 Driven Rod. The resistance to earth of the vertical rod in homogeneous earth can be developed by approximating the rod as a series of buried spherical elements (2-3). When the contributions of the elemental spheres are integrated along the length of rod and its image, the resistance to earth of the vertical rod is computed to be:

$$R_o = \frac{\rho}{2\pi l} \ln \frac{4l}{d} \quad (2-15)$$

where

$d$  = rod diameter, in cm,

$\rho$  = earth resistivity in ohm-cm,

$l$  = rod length, in cm.

An inaccuracy in the derived result arises from the assumption that equal incremental currents flow from the incremental spheres. Actually, more current per unit length flows into the soil near the earth's surface than at the lower end of the rod. It has been found empirically that the expression

$$R_o = \frac{\rho}{2\pi l} \ln \frac{3l}{d} \quad (2-16)$$

$$= \frac{0.159\rho}{l} \ln \frac{3l}{d}$$

$$= (2.306) \frac{0.159\rho}{l} \log \frac{3l}{d}$$

$$= \frac{0.366\rho}{l} \log \frac{3l}{d}$$

is a better approximation to the resistance to ground for a driven vertical rod. The net difference in resistance as given by Equations 2-15 and 2-16 is about 10 percent.

The resistance of the rod is directly affected by changes in the length of the rod and by the logarithm of the length. Changes in the diameter only show up as slight changes in the logarithm in Equation 2-15 and 2-16. Figures 2-6 and 2-7 show the measured changes in resistance that occurs with rod length and rod diameter. It is evident that effects of rod length do predominate over the effects of rod diameter.

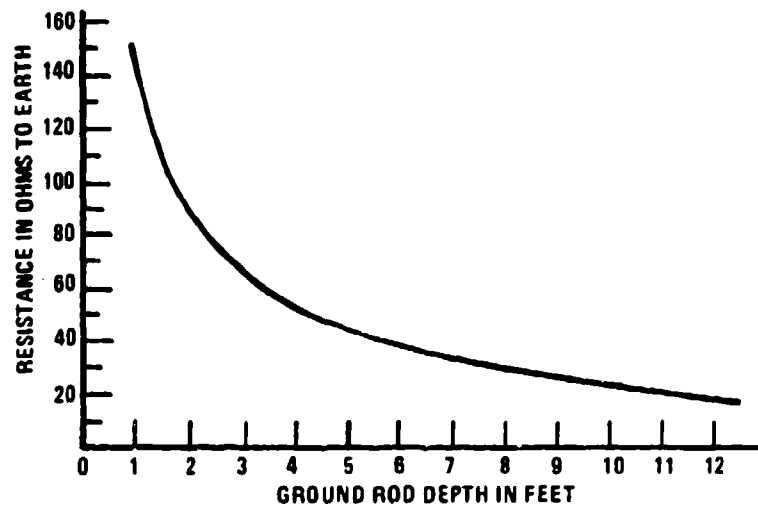


Figure 2-6. Effect on Rod Length Upon Resistance. (2-6)

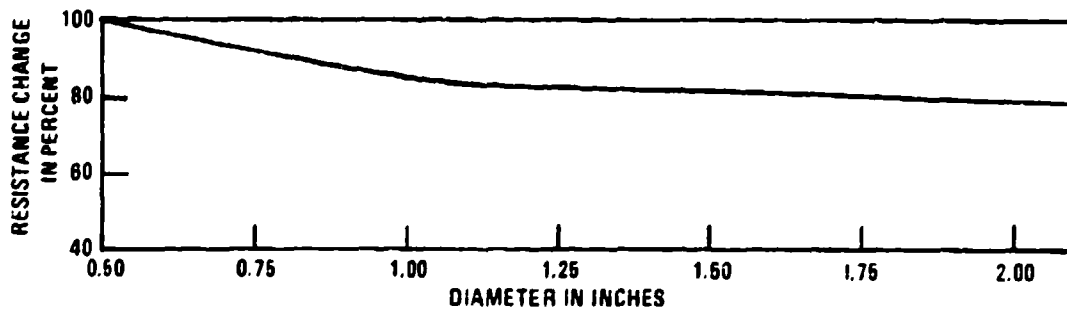


Figure 2-7. Effect of Rod Diameter Upon Resistance (2-6)

The earth surrounding the rod can be depicted conveniently as consisting of shells of earth of uniform thickness, as shown in Figure 2-8. The incremental resistance (in the direction of current flow) of each shell is given by

$$\frac{dR}{dr} = \rho \frac{d}{dr} \left( \frac{l}{A} \right) \quad (2-17)$$



which is a special form of Equation 2-1. The soil resistivity is  $\rho$  and  $dr$  is the incremental path length in the direction of current flow. The shell of earth nearest the electrode has the smallest area and thus exhibits the highest incremental resistance. This fact has two practical ramifications. First, lowest earth resistance is obtained with electrode configurations which have largest areas in contact with the earth. Second, changes which occur in the soil adjacent to the conductor have a significant effect on the electrode-to-earth contact resistance. For example, lightning discharge currents may heat the soil adjacent to the conductors, drying the soil or converting it to slag and thus increasing the electrode resistance to earth. One reason for providing a large contact area between the electrode and the earth is to minimize the current density in the soil immediately adjacent to the electrode, thus reducing the heating of the soil.

The current which flows into the ground rod flows outward through each equipotential shell, and the potential on the earth's surface at a distance,  $x$ , from the rod is (2-3)

$$E_x = \frac{0.366\rho I}{\ell} \log \left[ \frac{\ell}{x} + \sqrt{1 + \left(\frac{x}{\ell}\right)^2} \right]. \quad (2-18)$$

The ratio  $E_x/I$  is equivalent to  $R_x$ , that portion of resistance-to-ground of the rod which lies between the point  $X$  and infinity:

$$R_x = \frac{E_x}{I} = \frac{0.366\rho}{\ell} \log \left[ \frac{\ell}{x} + \sqrt{1 + \left(\frac{x}{\ell}\right)^2} \right]. \quad (2-19)$$

The ratio of  $R_x$  to  $R_o$  is

$$\frac{R_x}{R_o} = \frac{\log \left[ \frac{\ell}{x} + \sqrt{1 + \left(\frac{x}{\ell}\right)^2} \right]}{\log \frac{3\ell}{d}}, \quad (2-20)$$

where  $\ell$ ,  $d$ , and  $x$  are in the same units.

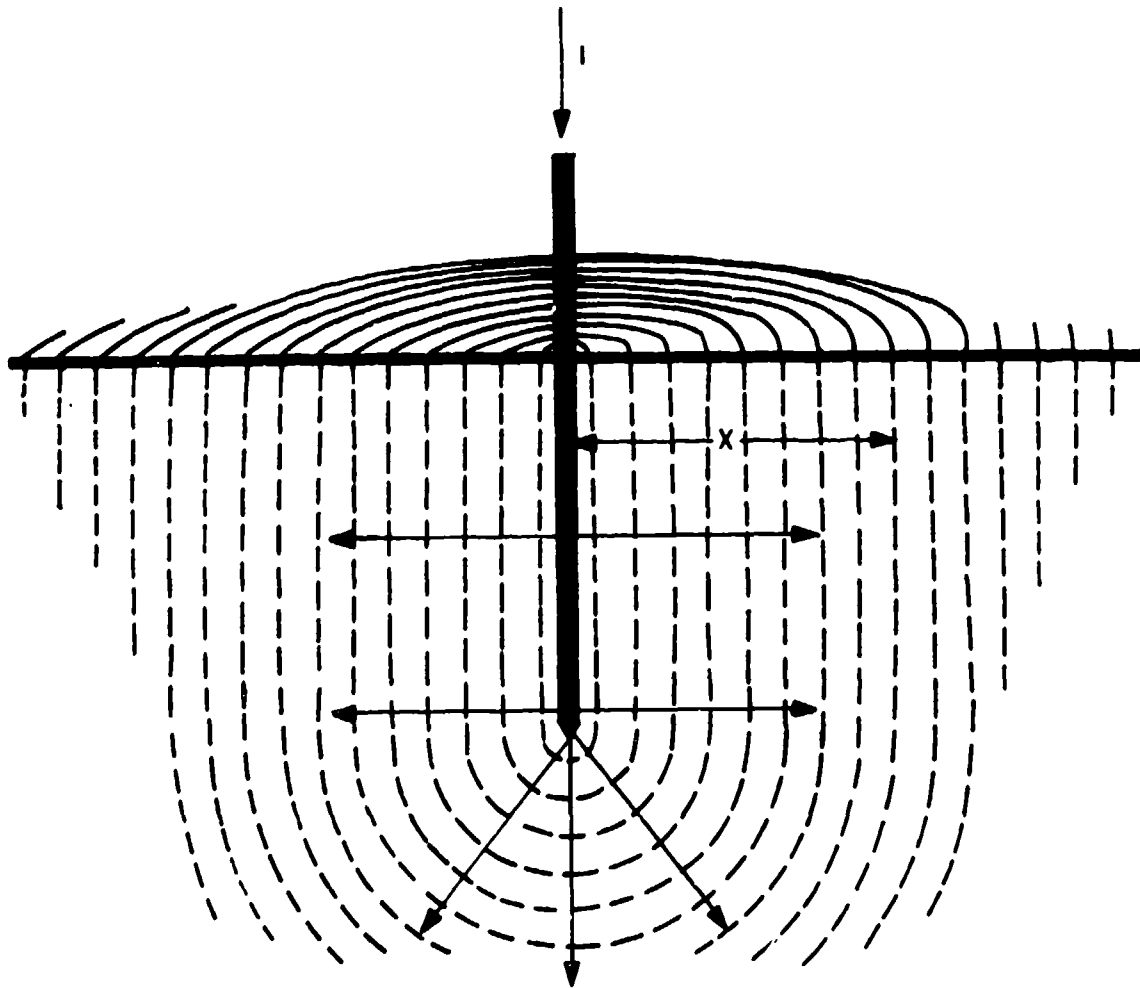


Figure 2-8. Earth Resistance Shells Surrounding a Vertical Earth Electrode

Equation 2-20 permits the area of influence of a single rod to be determined. For example, consider a 10-foot long, 1-inch diameter rod at distance  $x = \ell$ :

$$\frac{R_{\ell}}{R_o} = \frac{\log (1 + \sqrt{2})}{\log 360}$$

$$= 0.15$$

The ratio of 0.15 indicates that 85 percent of the total resistance to earth of a 10-foot long ground rod is established within 10 feet of the rod. For a 100-foot rod, 89 percent of the grounding resistance is obtained within 100 feet of the rod.

At a distance equal to two ground rod lengths,  $x = 2l$

$$\frac{R_{2l}}{R_o} = \frac{\log (0.5 + \sqrt{1.25})}{\log 360}$$

$$= 0.08$$

Thus 92 percent of the resistance of a 10-foot by 1-inch rod is obtained in a 20-foot radius cylinder. Similarly, 94 percent of the resistance of a 100-foot by 1-inch rod is obtained in a 200-foot radius cylinder. The resistance distribution for representative vertical electrodes is tabulated in Table 2-4.

Table 2-4

## Resistance Distribution for Vertical Electrodes

<u>Type of Rod Electrode</u>	<u>Total Resistance</u> (%)	<u>Approximate</u> <u>Distance from Rod</u> (feet)
3/4-inch pipe,	90	6
driven 3-feet	95	12
deep	98	31
	99	61
3/4-inch pipe,	90	9
driven 5-feet	95	18
deep	98	46
	99	92
1-1/4-inch pipe,	90	18
driven 10-feet	95	35
deep	98	88
	99	176
2-1/2-inch pipe,	90	25
driven 20-feet	95	69
deep	98	173
	99	345

Table 2-5

## Simple Isolated Electrodes

Order	Type	Conditions for Ranking	Resistance to Earth
$R_1$	Vertical-Rod		$R_1 = 0.366 \frac{\rho}{\ell} \log \frac{3\ell}{d}$
$R_2$	Buried Sphere	$\ell > 16d, r_s = \frac{\sqrt{d\ell}}{2}$	$R_2 = 0.08 \frac{\rho}{r} \left(1 + \frac{r}{2h}\right)$
$R_3$	Buried Circular Plate	$h > 3.2r_p = 4.5r_s$	$R_3 = 0.125 \frac{\rho}{r} \left(1 + \frac{r}{2.5h + r}\right)$
$R_4$	Half Buried Sphere (Buried Hemisphere)	$h > 1.1r_s = r_p$	$R_4 = 0.16 \frac{\rho}{r}$
$R_5$	Circular Plate on Surface of Earth		$R_5 = 0.25 \frac{\rho}{r}$
$R_6$	Buried Straight Rod or Wire	$h < 0.4\ell, r_p = \sqrt{d\ell}$	$R_6 = 0.366 \frac{\rho}{\ell} \left(\log \frac{\ell}{d} + \log \frac{\ell}{4h} + 0.34\right)$
$R_7$	Buried Right Angle	$h > 0$	$R_7 = 0.366 \frac{\rho}{\ell} \left(\log \frac{\ell}{d} + \log \frac{\ell}{4h} + 0.5 + 0.38 \frac{h}{\ell}\right)$
$R_8$	Buried Circle of Wire	$h < 0.8\ell$	$R_8 = 0.366 \frac{\rho}{\ell} \left(\log \frac{\ell}{d} + \log \frac{\ell}{4h} + 0.81\right)$
$R_9$	Buried Four Point Star	$h < 0.12\ell$	$R_9 = 0.366 \frac{\rho}{\ell} \left(\log \frac{\ell}{d} + \log \frac{\ell}{4h} + 1.26 - 3.72 \frac{h}{\ell}\right)$

Note:  $\log X = \log_{10} X$ ,  $r_s$  = radius of sphere,  $r_p$  = radius of plate;  $\ell$  = total length of buried rod or wire, in cm;  $d$  = diameter of rod or wire, in cm; and  $h$  = depth of burial, in cm.  $\rho$  = ohm-cm. The ranking assumes that each electrode has equal surface area in contact with the earth.

**2.6.1.2 Other Commonly Used Electrodes.** Table 2-5 lists a number of simple isolated earthing electrodes along with approximate formulas for their resistance to earth. The plate and spherical electrodes are extensive in area, whereas the vertical rod, the horizontal rod (or wire), the star, and the circle are extensive in length. The electrodes in Table 2-5 have been ranked after being normalized for equal surface area in contact with the earth. The order of ranking is such that the lowest resistance-to-earth electrode (the most effective) heads the list. As an example, a circular plate lying on the earth's surface is a more effective electrode (has a lower resistance to earth) than a buried, horizontal rod which has the same area in contact with the earth, assuming that the rod is buried at a depth less than 40 percent of its length.

The resistance to earth provided by horizontal conductors as a function of length is shown in Figure 2-9 for two depths of burial. Note that as the length is doubled, the resistance is approximately halved. The curves of Figure 2-9 assume that the conductors are laid out in a straight line. If the strips are coiled or curved, the resistance tends to be higher because the cross-sectional area of the soil affected is less.

The resistance of a plate ground is dependent upon the area of the plate. The variation of resistance as a function of the radius of a circular plate is illustrated in Figure 2-10 for three depths of burial. These curves are calculated for a plate in soil of uniform resistivity of 10,000 ohm-cm. Similar relationships hold for rectangular plates; the curves as shown should be considered to indicate the behavior of resistance as a function of area rather than as a prediction of the resistance of plate of a given area.

**2.6.2 Resistance of Multiple Electrodes.** The theoretical resistance of an electrode, such as given by Equation 2-16, is obtained only at an infinite distance from the electrode. As shown in Section 2.6.1.1, however, most of the resistance of a single electrode is obtained within a reasonable distance from the electrode. (For a vertical rod, better than 90 percent is realized within two rod lengths.) If two or more electrodes are closely spaced, however, the total effective resistance of neither is realized. This interaction prevents the resistance of  $N$  electrodes connected in parallel from being  $1/N$  times the resistance of one of the electrodes. For this reason, the crowding of multiple vertical rods is not as beneficial in terms of dollar cost per ohm as is achievable with fewer rods properly spaced. If the electrodes in a multiple electrode installation are separated by adequate distances, the interactive influence is minimized. The separation between driven vertical ground rods in a group of rods should not be less than the length or greater than twice the length of an individual rod.

**2.6.2.1 Two Vertical Rods in Parallel.** Expressions for the resistance of multiple electrodes are more complex than those for isolated electrodes. To illustrate, consider two rods driven into the earth with their tops flush with the surface as shown in Figure 2-11. The two rods are electrically in parallel, but the presence of one rod affects the resistance of the other. The resistance-to-earth of two rods (2-9) is

$$R = \frac{\rho}{4\pi\ell} \left[ \ln \left( \frac{2\ell + \sqrt{s^2 + 4\ell^2}}{s} \right) + \frac{s}{2\ell} - \sqrt{\frac{s^2 + 4\ell^2}{2\ell}} + \ln \frac{4\ell}{r} - 1 \right], \quad (2-21)$$

where  $s$  = spacing between rods.

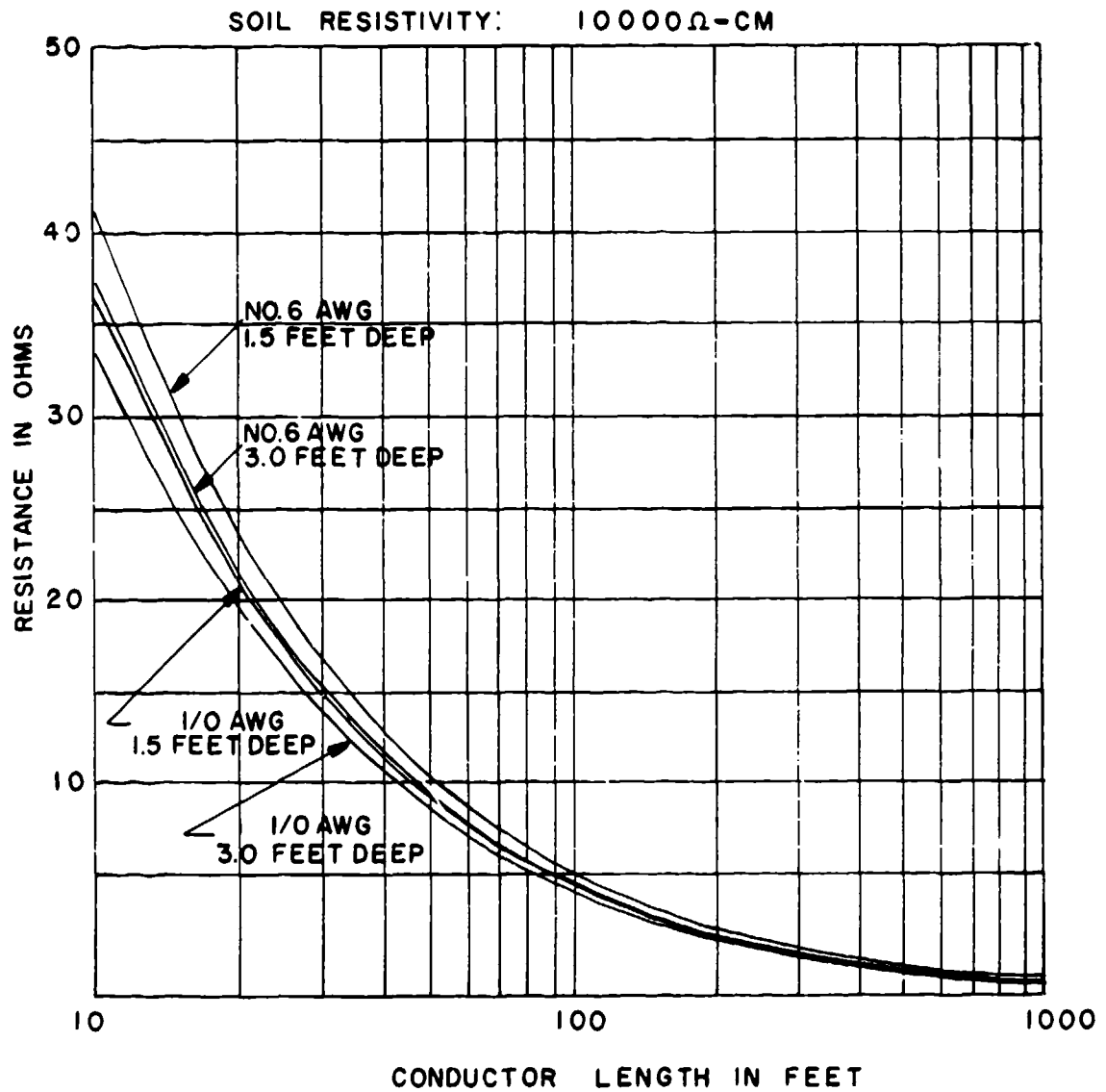


Figure 2-9. Resistance of Buried Horizontal Conductors

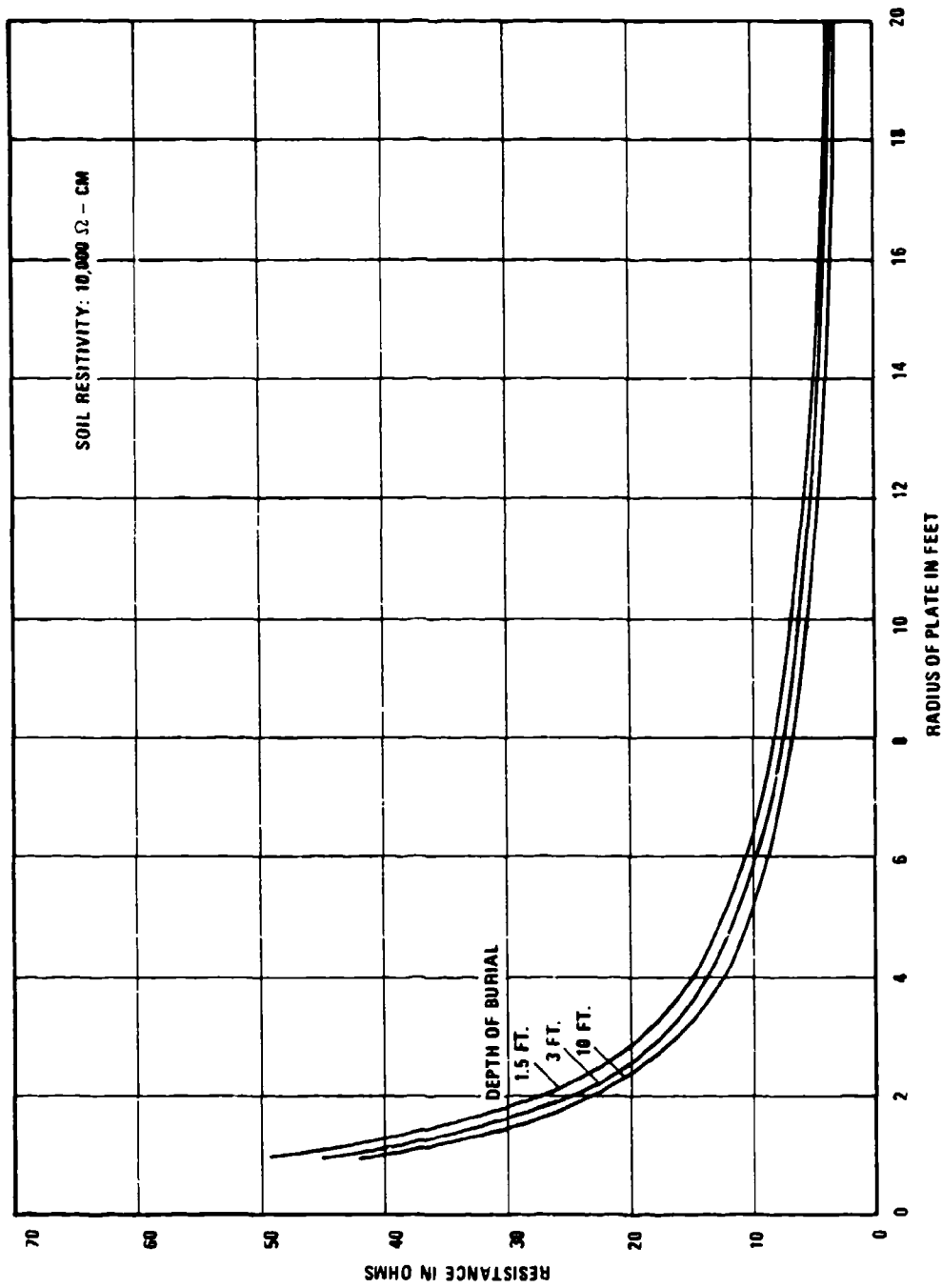


Figure 2-10. Resistance of Buried Circular Plates

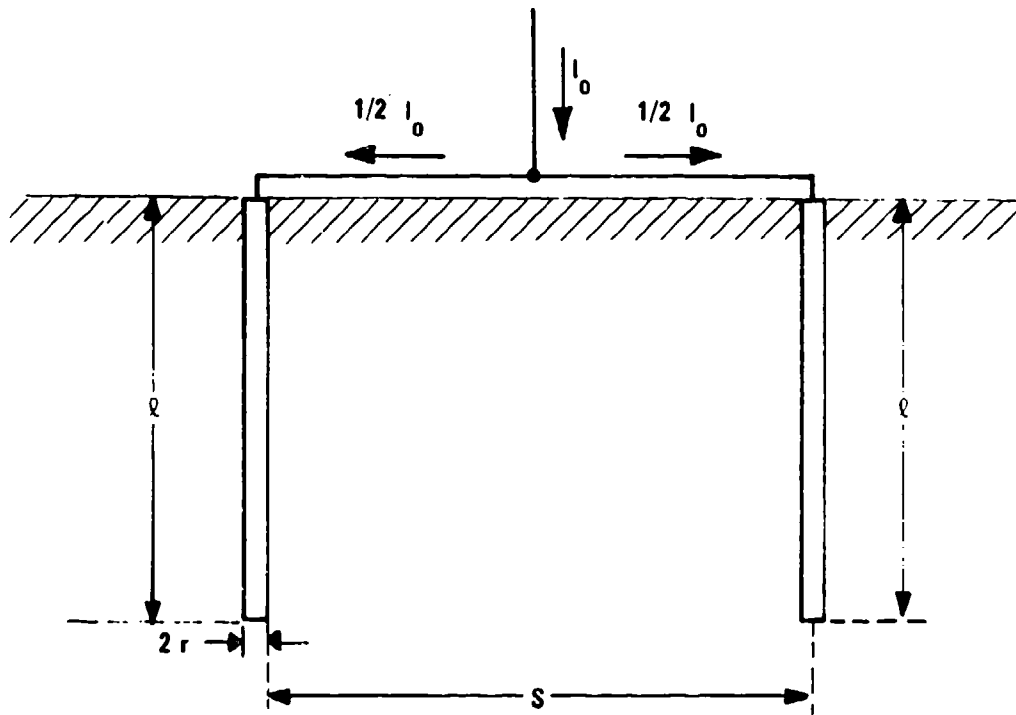


Figure 2-11. Ground Rods in Parallel



For the condition of  $s > \ell$ ,

$$R = \frac{\rho}{4\pi\ell} \left( \ln \frac{4\ell}{r} - 1 \right) + \frac{\rho}{4\pi s} \left( 1 - \frac{\ell^2}{3s^2} + \frac{2}{5} \frac{\ell^4}{s^4} + \dots \right). \quad (2-22)$$

For  $s < \ell$ ,

$$R = \frac{\rho}{4\pi\ell} \left( \ln \frac{4\ell}{r} - 2 + \ln \frac{4\ell}{s} + \frac{s}{2\ell} - \frac{s^2}{16\ell} 2 + \dots \right). \quad (2-23)$$

For  $s = \ell$ ,

$$R = \frac{\rho}{4\pi\ell} \left( \ln \frac{4\ell}{r} - 0.18 \right). \quad (2-24)$$

If a number,  $N$ , of equal length vertical ground rods (with tops flush with the surface) are separated equally along a straight line and connected together by an insulated conductor at the tops of the rods, the resultant resistance will be somewhat greater than  $1/N$  times the resistance of single isolated rod. For  $N$  rods of length  $\ell$  at spacing  $s$ , the total resistance  $R_N$  is given by

$$R_N = \frac{1}{N} \frac{\rho}{2\pi\ell} \left[ \ln \frac{4\ell}{r} - 1 + \frac{2\ell}{s} \ln \frac{2N}{\pi} \right] \quad (2-25)$$

where  $r$  is the radius of each rod.

#### 2.6.2.2 Square Array of Vertical Rods.

The resistance of a square array of rods is

$$\begin{aligned} R_r &= \frac{\text{Resistance of one rod}}{\text{Number of rods in array, } N} \times \text{Resistance ratio, } K \\ &= \frac{R_{\text{one rod}}}{N} K. \end{aligned} \quad (2-26)$$

Figure 2-12 shows the value of  $K$  for a square array of  $N$  equally spaced, equal length rods at spacings up to 10 times a rod length. The distance from a rod to its closest neighbor in the array is  $s$ , and the various curves in Figure 2-12 correspond to values of  $s$ , stated as integral multiples of rod length. To illustrate the use of Figure 2-12, consider a 5 by 5 array of 25 rods, each spaced one length from its closest neighbor. From the  $s = \ell$  curve, it is found that the resistance ratio is 2.8 for a 25-rod group. The parallel resistance of the 25 rods is therefore 2.8 times one twenty-fifth ( $1/N$ ) of the resistance that one of the rods would exhibit if isolated.

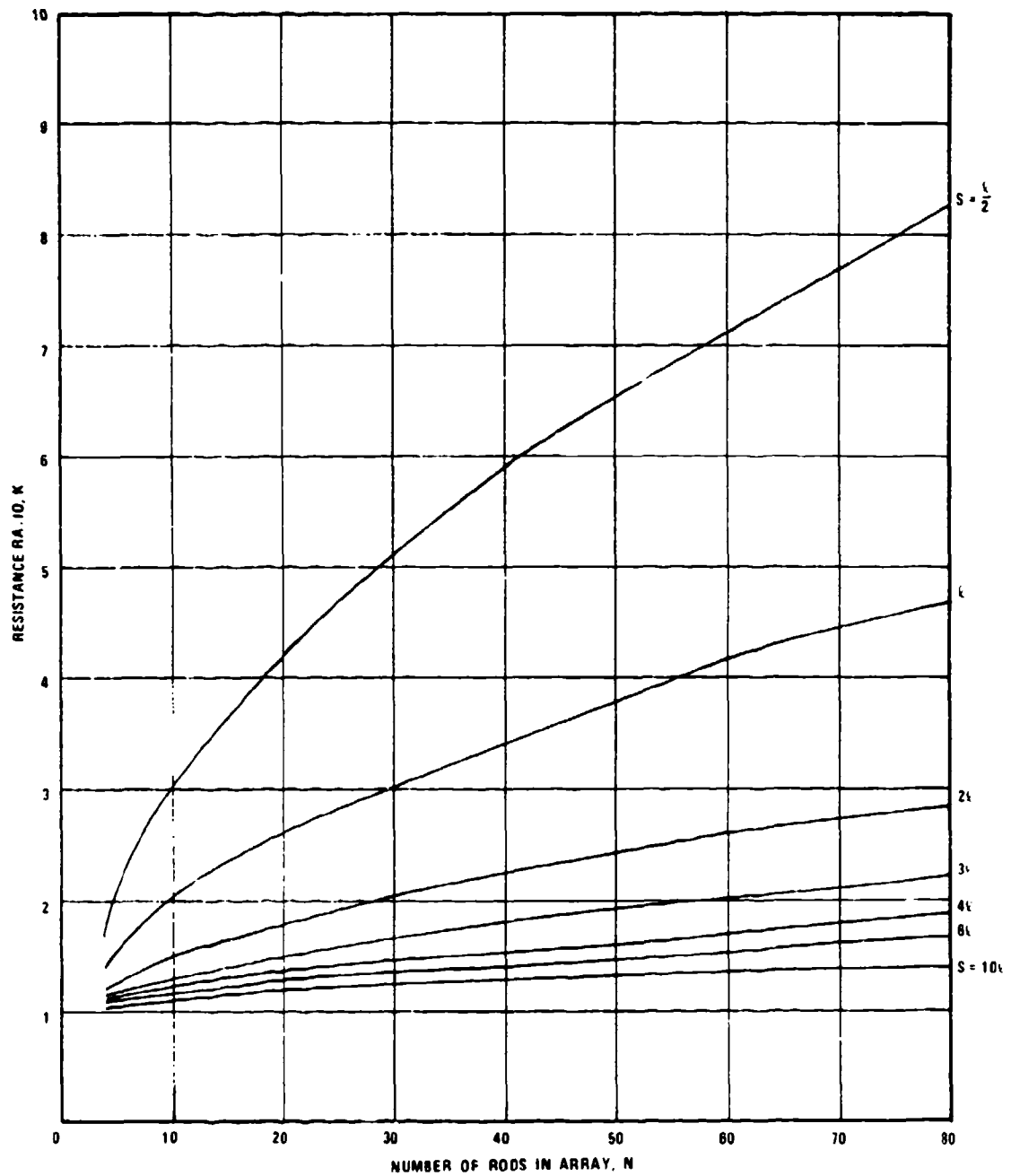


Figure 2-12. Ratio of the Actual Resistance of a Rod Array to the Ideal Resistance of  $N$  Rods in Parallel

**2.6.2.3 Horizontal Grid (Mesh).** Earth electrode subsystems for electric power stations and substations must be designed both to provide low resistance to earth and to minimize voltage gradients at the earth's surface (see Section 2.8.1). A common electrode design for such applications is a grid, or mesh, of horizontal rods or wires connected at each crossing. The resistance to earth for a square or a rectangular grid can be calculated from the following Equation (2-3):

$$R_w = \rho \left( \frac{1}{2D_e} + \frac{1}{L_{tot}} \right) \quad (2-27)$$

where

$\rho$  = earth resistivity,

$L_{tot}$  = total length of conductors used,

$\frac{\pi D_e^2}{4} = A$  = area covered by grid, and

$D_e$  = effective diameter of grid.

As an example, consider a square grid that has dimensions of 30.5 m x 30.5 m (100 feet by 100 feet) with conductors spaced 3.05 m (10 feet) apart. Thus there are 100 meshes with a total conductor length of 670 m (2200 feet). The area of the array is 929 square meters (10,000 square feet) with an effective diameter of

$$\begin{aligned} D_e &= \sqrt{\frac{4A}{\pi}} \\ &= 113 \text{ feet} \\ &= 3440 \text{ cm} \end{aligned}$$

Thus the resistance to earth, by Equation 2-27, is

$$\begin{aligned} R &= \frac{\rho}{2(3440)} + \frac{\rho}{67100} \\ &= 1.45 \times 10^{-4} \rho + 0.15 \times 10^{-4} \rho \\ &= 1.6 \times 10^{-4} \rho \text{ ohms} \end{aligned} \quad (2-28)$$

2.6.2.4 Vertical Rods Connected by a Grid. The resistance of a bed of vertical rods, interconnected with a wire grid is (2-10)

$$R_t = \frac{R_\omega R_r - R_m^2}{R_\omega + R_r - 2R_m} \quad (2-29)$$

where

$R_\omega$  = resistance of wire grid as given by Equation 2-27

$L_{tot}$  = length of conductors in grid

$R_r$  = resistance of bed of rods, as found from Figure 2-12

$R_m$  = mutual resistance which accounts for interaction of rods on grid

$$R_m = \frac{0.73}{L_{tot}} \rho \left[ \log \frac{2L_{tot}}{\sqrt{2 r_g h}} - 1 \right], \text{ or} \quad (2-30)$$

$$= \frac{0.73}{L_{tot}} \rho \log \frac{2L_{tot}}{l} \quad (2-31)$$

where

$r_g$  = radius of grid wire,

$h$  = depth of grid, if buried, and

$l$  = length of rod, if the grid is near surface.

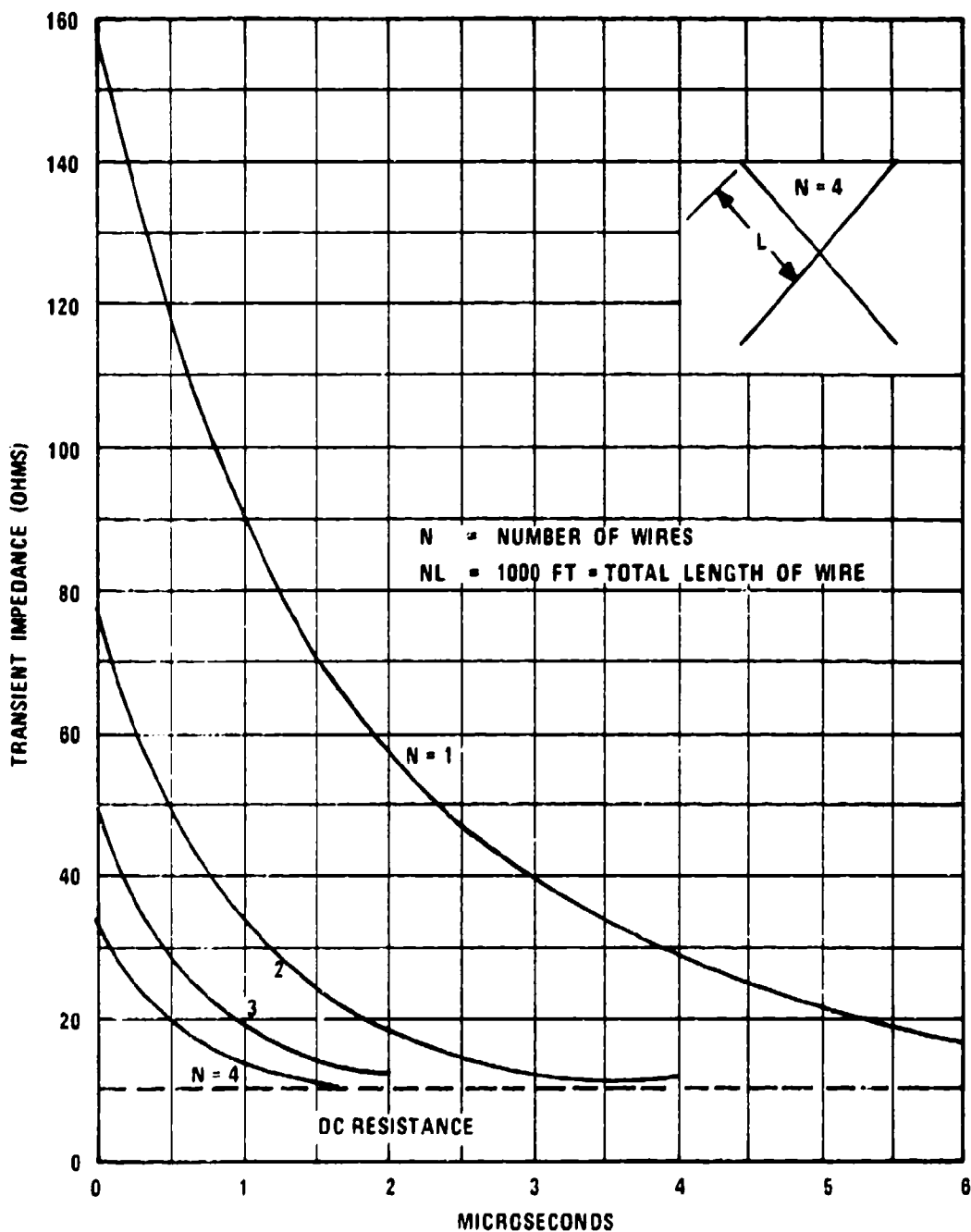


Figure 2-13. Transient Impedance of an Earth Electrode Subsystem as a Function of the Number of Radial Wires

**2.6.3 Transient Impedance of Electrodes.** The expressions given for electrode resistance assume perfect conductivity for the conductors of an electrode. Such an assumption introduces very little error in the calculation of the electrode dc resistance, but if the electrode must dissipate the impulsive energy of a lightning stroke, its impedance as a function of time must be considered. When a single star electrode, containing 305 meters (1000 feet) of conductor, is subjected to a surge of lightning current, the initial value of its effective impedance is about ten times the dc resistance (2-11). This initial value is termed the surge impedance. As the wave of energy propagates through the electrode system, more and more of the wire of the electrode makes effective contact between the propagating energy and the medium which dissipates the energy. It is clear that a given length of wire will couple lightning energy more efficiently into the earth if the electrode is in the form of a star than if it were a single conductor. This is illustrated in Figure 2-13 where it is indicated that as the energy surges down an electrode (at a velocity in the neighborhood of 100 meters (333 feet) per microsecond), the transient impedance of the electrode decreases and approaches the dc resistance value.

**2.6.4 Effects of Nonhomogeneous (Layered) Earth.** The previous derivations assumed homogeneous earth. A qualitative understanding of the effects of non-uniform earth resistivity can be deduced from Figure 2-14 which illustrates the electric equipotential surfaces and current flow in layered earth when the earthing electrode is a small hemisphere. The lines radiating outward from the earth electrode indicate the flow of current. Not surprisingly, if the resistivity of the deeper layer is high, relative to the upper layer, nearly all of the current is confined to the upper layer of earth.

**2.6.4.1 Hemispherical Electrode.** An approximate expression (2-3) for the resistance to earth of a small hemispherical electrode in layered earth is

$$R = \frac{\rho_1}{2\pi r} + 0.366 \frac{\rho_1}{h} \log \frac{\rho_1 + \rho_2}{2\rho_1} \quad (2-32)$$

where

$r$  = hemisphere radius (assumed less than  $h$ ),

$h$  = thickness of superficial layer,

$\rho_1$  = resistivity of superficial layer,

$\rho_2$  = resistivity of deep layer.

An interesting example is the case of a superficial layer of low resistivity soil ( $\rho = 10^3$  ohm-cm) over granitic ( $\rho = 10^6$  ohm-cm):

$$\begin{aligned} R &= \frac{10^3}{8.3r} + \frac{366}{h} \log \left( \frac{1}{2} + 500 \right), \\ &= \frac{160}{r} \left[ 1 + \frac{6.2}{h/r} \right], \end{aligned} \quad (2-33)$$

where  $r$  and  $h$  are measured in centimeters. If  $h < 6.2 r$ , the resistance to earth will be greatly influenced by the resistivity of the granite underlayment; if  $h > 6.2 r$ , the resistance approaches that for homogeneous earth with resistivity,  $\rho_1$ .

#### 2.6.4.2 Vertical Rod.

When a vertical rod is driven through a high resistivity superficial (upper) layer into a lower resistivity subsoil, an adjustment can be made to the resistance to earth expression for homogeneous soil by substituting a reduced "effective length" of the ground rod. Letting  $\ell'$  be the effective length (2-3)

$$\ell' = \ell - h \left( 1 - \frac{\rho_2}{\rho_1} \right) \quad (2-34)$$

where

$\ell$  = physical length of rod,

$\rho_1$  = resistivity of upper layer,

$\rho_2$  = resistivity of subsoil, and

$h$  = depth of upper layer.

Note that if  $\rho_1 \gg \rho_2$ , the effective length of the rod is reduced to  $\ell - h$ . When the subsoil has a higher resistivity than the top layer of soil ( $\rho_2 > \rho_1$ ), the current discharged through a slender vertical rod with length equal to the thickness of the superficial layer of soil will tend to remain in the superficial layer of soil. The "mean path" of the superficial layer current, that is the radial distance at which half the discharge current has entered the deeper soil, is approximately (2-3)

$$X = \frac{\rho_2}{\rho_1} h \quad (2-35)$$

If the dimensions of the earth electrode subsystem are large compared to the thickness of the upper stratum, the upper layer becomes insignificant and the resistance to earth can be computed as though the soil were homogeneous with resistivity equal to  $\rho_2$ , the resistivity of the subsoil.

#### 2.6.4.3 Grids.

A useful approximation for the resistance-to-earth of a horizontally extensive electrode system is given by Equation 2-27.

If the soil has a superficial layer with resistivity  $\rho_1$  and a subsoil with resistivity  $\rho_2$ , the resistance to earth of a grid in the superficial layer is given by (2-3)

$$R \approx \frac{\rho_2}{2D_e} + \frac{\rho_1}{L_{tot}} \quad (2-36)$$

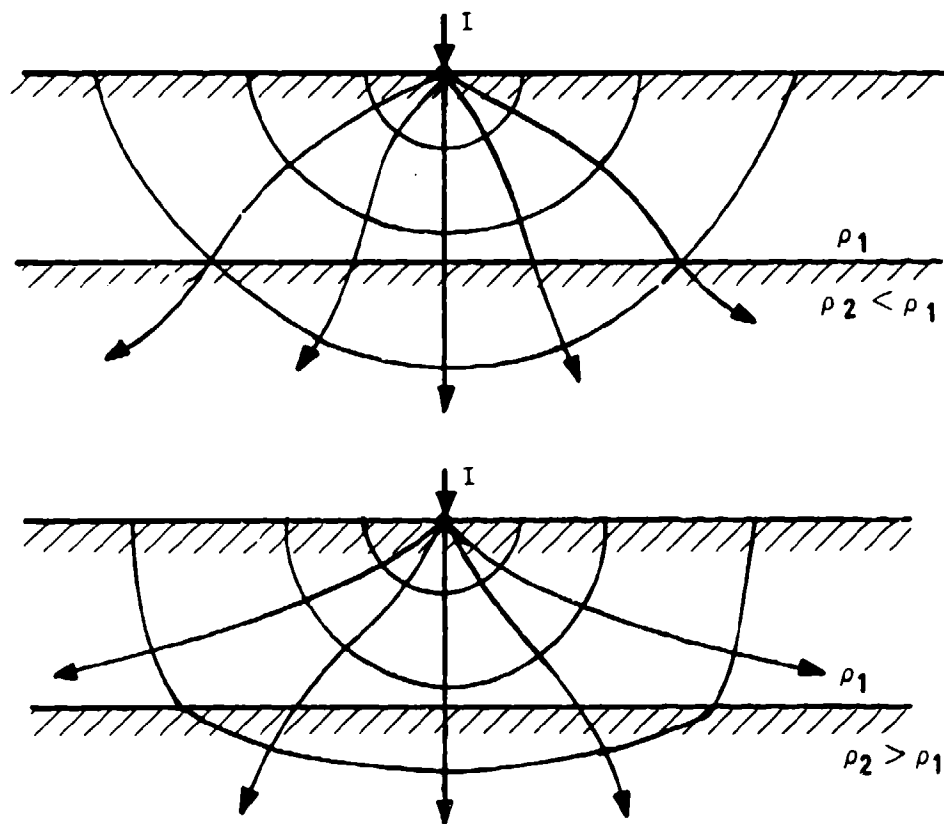


Figure 2-14. Current Distribution in Nonuniform Soil



when  $\rho_2 \gg \rho_1 \left( \frac{2D_e}{L_{tot}} \right)$ , the earthing resistance is approximately

$$R \approx \frac{\rho_2}{2D_e} \quad (2-37)$$

and when  $\rho_1 \gg \rho_2 \left( \frac{L_{tot}}{2D_e} \right)$ , it is approximately

$$R \approx \frac{\rho_1}{L_{tot}} \quad (2-38)$$

If, for example, the diameter,  $D_e$ , of the grid equals 500 meters, the resistivity,  $\rho_1$ , of the superficial layer equals 10,000 ohm-meters, the resistivity,  $\rho_2$ , of the subsoil equals 200 ohm-meters, and the length,  $L_{tot}$ , of the conductors in the grid equals 4,000 meters, then

$$R \approx 2.7 \text{ ohms}$$

Burying the grid within the lower resistivity subsoil would reduce the resistance-to-earth to about 0.4 ohms. Conversely, if the  $\rho_2 = 10,000$  ohm-meters, and  $\rho_1 = 200$  ohm-meters, then

$$R \approx 10 \text{ ohms}$$

regardless of the depth of the grid.

## 2.7 MEASUREMENT OF RESISTANCE-TO-EARTH OF ELECTRODES.

**2.7.1 Introduction.** The calculated resistance of a given electrode system is based on a variety of assumptions and approximations that may or may not be met in the final installation. Because of unexpected and uncontrolled conditions which may arise during construction, or develop afterward, the resistance of the installed electrode must be measured to see if the design criteria are met. In an existing facility, the resistance of the electrode system must be measured to see if modifications or upgrading is necessary. Two commonly used methods for measuring the resistance to earth of an electrode are the triangulation method and the fall-of-potential method.

**2.7.2 Fall-of-Potential Method.** This technique involves the passing of a known current between the electrode under test and a current probe,  $C_2$ , as shown in Figure 2-15(a). The drop in voltage between the earth electrode and the potential electrode,  $P_2$ , located between the current electrodes is then measured; the ratio of the voltage drop to the known current gives a measure of the resistance. (By using a voltage measuring device - a null instrument or one having a high impedance - the contact resistance of the potential electrode will have no appreciable effect on the accuracy of the measurement.) Several resistance measurements are taken by moving the potential probe,  $P_2$ , from the position of the earth electrode, along a straight line to the current probe,  $C_2$ , which is left in position. The data obtained is then plotted as resistance versus distance from the earth electrode as illustrated in Figure 2-15(b). This is the test method recommended for measurement of single rod or multi-rod earth electrode subsystems.

**2.7.2.1 Probe Spacing.** Current flow into the earth (see Figure 2-8) surrounding an electrode produces shells of equipotential around the electrode. A family of equipotential shells exists around both the electrode under test and the current reference probe,  $C_2$ . The sphere of influence of these shells is proportional to the size of each respective electrode. (See, for example, Section 2.6.1.1.) The potential probe,  $P_2$ , in Figure 2-15 provides an indication of the net voltage developed at the earth's surface by the combined effect of these two families of shells. If the electrode under test and the current reference probe are so close that their equipotential shells overlap, the surface voltage variation as measured by  $P_2$  will vary as shown in Figure 2-16(a). Since the current flowing between the electrodes is constant for each voltage measurement, the resistance curve will have the same shape as the voltage curve. For close electrode spacings, the continuously varying resistance curve does not permit an accurate determination of resistance to be made.

By locating the current reference probe,  $C_2$ , far enough away from the electrode under test to ensure that the families of equipotential shells do not overlap, a voltage curve like that shown in Figure 2-16(b) will be obtained to produce the type of resistance curve shown in Figure 2-15.

When the distance,  $D$ , between the electrode under test and the current reference probe is very large compared to the dimensions of the earth electrode subsystem under test, the latter can be approximated as a hemisphere and interaction between the two electrodes is negligible. When these assumptions are met, the potential at a point at distance  $x$  from the electrode under test is:

$$U_x = \frac{\rho I}{2\pi x} - \frac{\rho I}{2\pi(D-x)} = \frac{\rho I}{2\pi} \left( \frac{1}{x} - \frac{1}{D-x} \right) \quad (2-39)$$

where  $\rho$  is the average soil resistivity; the minus sign indicates that the current,  $I$ , flows into  $C_1$ , and out from  $C_2$ .

Assume that the electrode under test is equivalent to a hemisphere with radius,  $r$ . At the surface of this hemisphere, the potential is found by letting  $x = r$ :

$$U_o = \frac{\rho I}{2\pi} \left( \frac{1}{r} - \frac{1}{D-r} \right) \quad (2-40)$$

The potential difference between  $C_1$  and  $P_2$  is the voltage that is being measured and is:

$$\begin{aligned} V_o^x &= (U_o - U_x) \\ &= \frac{\rho I}{2\pi} \left( \frac{1}{r} - \frac{1}{D-r} - \frac{1}{x} + \frac{1}{D-x} \right) \end{aligned} \quad (2-41)$$

when  $x = r$

$$V_o^x = 0$$

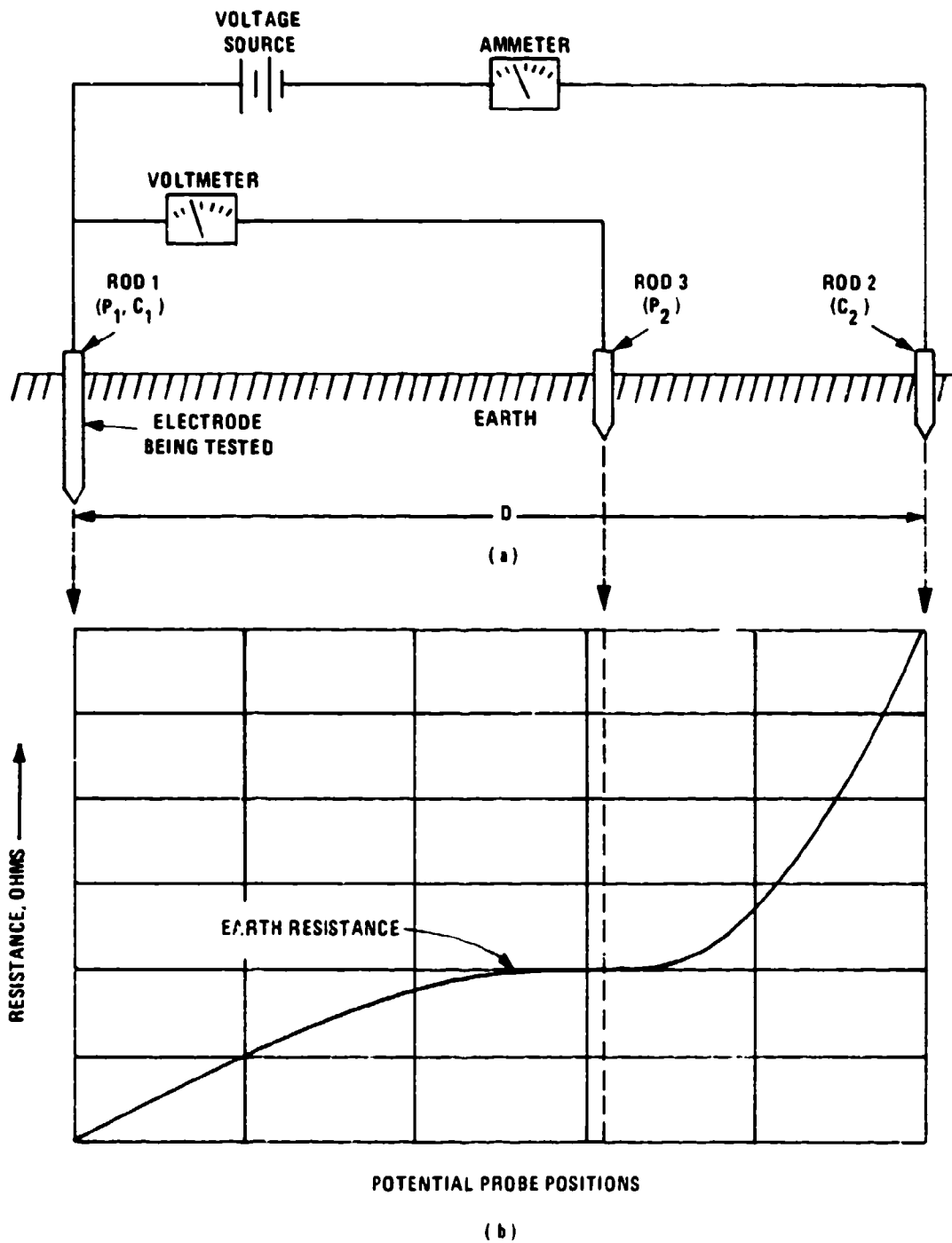
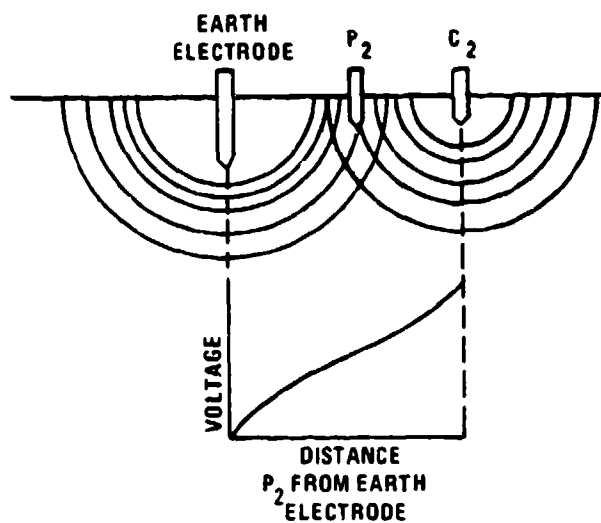
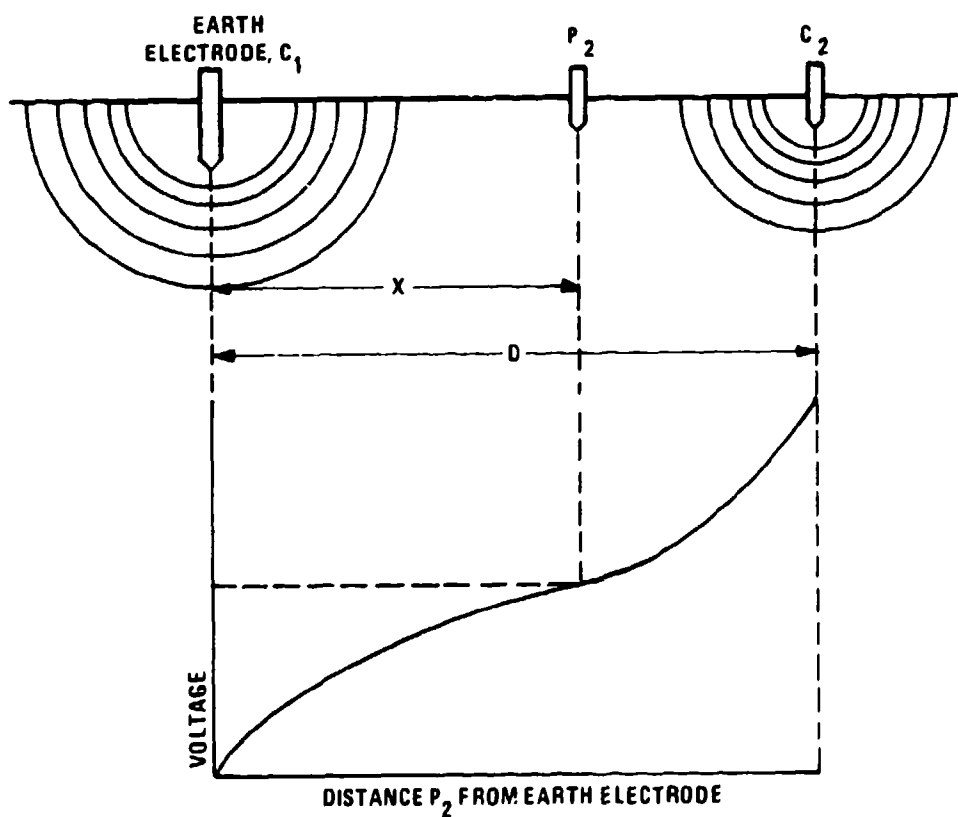


Figure 2-15. Fall-of-Potential Method for Measuring the Resistance of Earth Electrodes



(a) CLOSELY SPACED ELECTRODES



(b) WIDELY SPACED ELECTRODES

Figure 2-16. Effect of Electrode Spacing on Voltage Measurement

If the  $r_2$  is the radius of the hemisphere that is equivalent to the current probe,  $C_2$ , and  $r$  is the equivalent radius of the electrode under test, it is seen that when  $x = D - r_2$

$$V_o^x = \frac{\rho I}{2\pi} \left( \frac{1}{r} - \frac{1}{D-r} - \frac{1}{D-r_2} + \frac{1}{r_2} \right) \quad (2-42)$$

If  $D \gg r_2$  or  $r$

$$V_o^x = \frac{\rho I}{2\pi} \left( \frac{1}{r} + \frac{1}{r_2} \right) \quad (2-43)$$

But the true value of resistance corresponds to

$$V_o^x = \frac{\rho I}{2\pi r} \quad (2-44)$$

which is found when  $0 < x < D - r_2$ .

In order for the measurement of  $V_o^x$  to yield the correct value of resistance to earth; it can be seen that the error term in Equation 2-41 must be zero, i.e.,

$$\frac{1}{D-x} - \frac{1}{x} - \frac{1}{D-r} = 0 \quad (2-45)$$

$$x(D-r) - (D-r)(D-x) - x(D-x) = 0$$

$$Dx \left( 1 - \frac{r}{D} \right) - D(D-x) \left( 1 - \frac{r}{D} \right) - x(D-x) = 0.$$

Again if  $D \gg r$

$$x^2 + Dx - D^2 = 0 \quad (2-46)$$

which can be solved as follows:

$$\begin{aligned}
 x &= -\frac{D}{2} + \frac{1}{2} \sqrt{D^2 + 4D^2} & (2-47) \\
 &= \frac{D}{2} (\sqrt{5} - 1) \\
 &= \frac{D}{2} (2.236 - 1) \\
 &= \frac{D}{2} (1.236) \\
 &= 0.618D
 \end{aligned}$$

Thus the true value of resistance to earth corresponds to the ratio of the potential difference to the measured current when  $x$  is 62 percent of the distance,  $D$ , from the electrode under test to the current probe,  $C_2$ . It is important to remember that  $D$  is measured from the center of the electrode under test to the center of the current probe and that  $D$  is large relative to the radius of the electrode under test.

Figure 2-17 shows an example of data taken with the fall-of-potential method. The correct resistance of 13 ohms corresponds to the potential probe location of 27.4 meters (90 feet) which is 62 percent of the distance to the current probe.

Resistance of the electrode under test with respect to infinity (the true definition of the resistance to earth) is

$$R = \frac{U_o}{I} = \frac{\rho}{2\pi} \left( \frac{1}{r} - \frac{1}{D-r} \right) \quad (2-48)$$

Thus any value of  $D$  less than infinity causes the measured resistance to be in error. The error can be estimated by observing that

$$R = \frac{\rho}{2\pi r} \left( 1 - \frac{r}{D-r} \right)$$

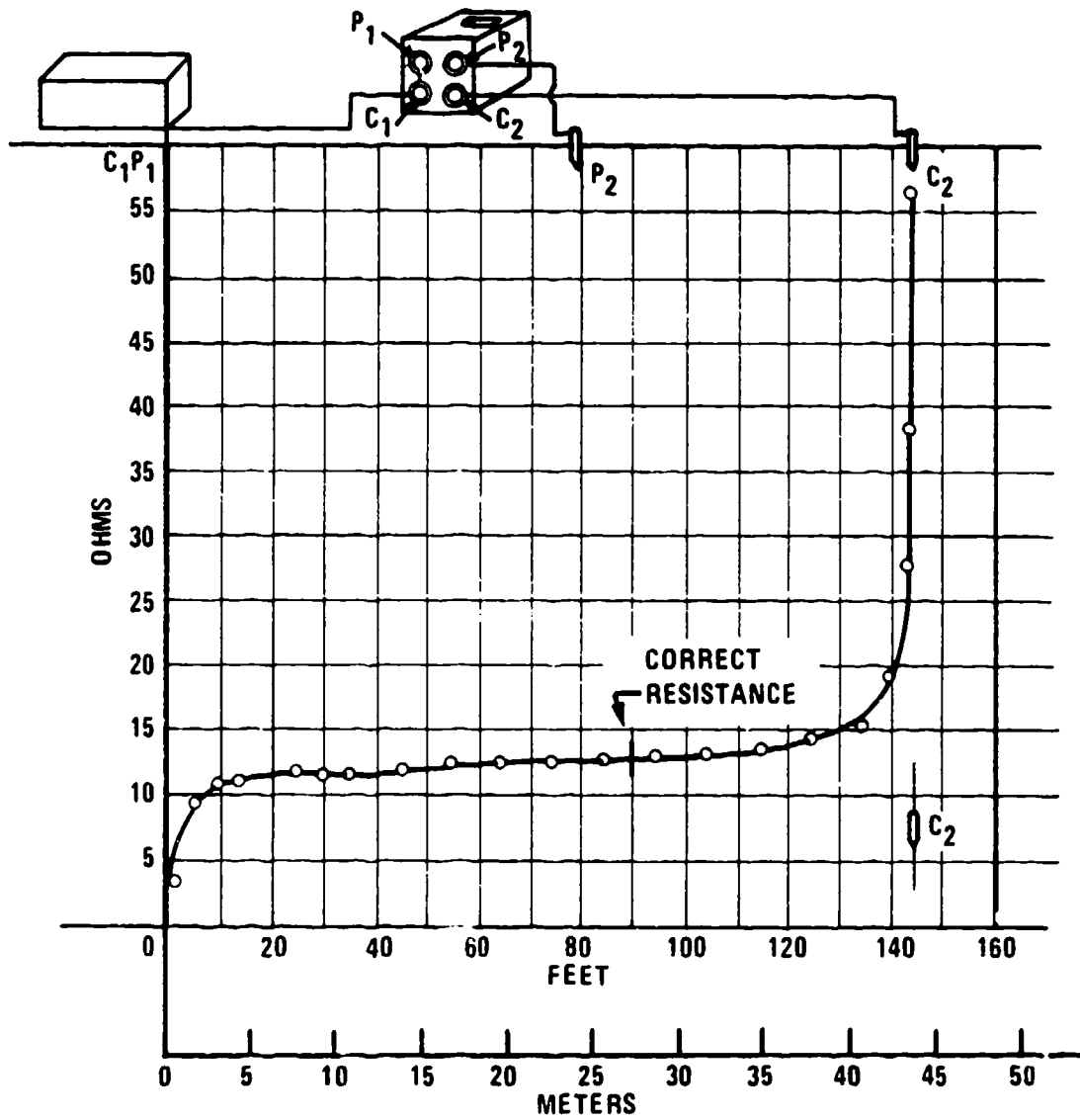


Figure 2-17. Resistance Variations as Function of Potential Probe Position in Fall-of-Potential Method (2-12)

Remembering that

$$R = \frac{\rho}{2\pi r}$$

is the true resistance, it is evident that if  $D = 5r$ , the error will be 25 percent, if  $D = 11r$ , the error is 10 percent; if  $D = 26r$ , the error is 4 percent, etc.

The equivalent radius of a large electrode system can be determined from

$$r_e = \sqrt{\frac{A}{\pi}} \quad (2-49)$$

where

$A$  = the area covered by the system.

Consider a rectangular grid 10 meters by 10 meters. Its effective radius is

$$r_e = \sqrt{\frac{10^2}{\pi}} = 5.64 \text{ meters.}$$

For an accuracy of 90 percent, the probe  $C_2$  should be positioned at

$$D = 11 \times 5.64$$

$$= 62 \text{ meters or } 203 \text{ feet away.}$$

A conservative estimate which leads to improved accuracy of the effective radius is that it is equal to one half the longest diagonal dimension ( $D_d$ ) of the array. Thus for an accuracy of 90 percent, the location for  $C_2$  should be

$$11 \times (0.5 D_d) \text{ or } 5.5 D_d,$$

which is the basis for the frequently quoted rule of thumb of 5 times the longest diagonal of the area of the electrode under test. Table 2-6 gives the percentage accuracies obtained at probe locations up to 50 times the longest diagonal.

**2.7.2.2 Extensive Electrode Subsystems (2-13).** When the earth electrode subsystem is extensive, it is frequently difficult to locate the current probe at a distance of even five times the largest dimension and measurements of resistance to earth are subject to large errors. In addition, a connection to the electrical center of the subsystem may not be possible. Figure 2-18 shows a set of resistance curves for an extensive earth electrode subsystem obtained at current probe spacings of up to 304 meters (1000 feet). Each curve corresponds to a particular distance,  $C_k$ , of the current probe from the point of connection to the earth electrode subsystem. The potential probe spacing,  $P$ , is the independent variable.



Table 2-6

Resistance Accuracy Versus Probe C<sub>2</sub> Spacing

Accuracy (percent)	Probe Spacing
90	5 x diagonal under test
95	10 x diagonal under test
98	25 x diagonal under test
99	50 x diagonal under test

On each curve the points corresponding to 62 percent of the distance to the current probe have been connected. It is evident that as the current probe location is moved farther out, the 62 percent value is decreasing. The true value of resistance can be estimated by extrapolating the connecting line to its asymptotic value. Because none of the curves in Figure 2-18 level out, even the largest spacing of the current probe is evidently too small for a direct reading of the resistance. Basic assumptions for the fall-of-potential measurement are that (1) the electrode to be measured can be approximated as a hemisphere and (2) the connection to the earth electrode is made at its electrical center. Since the location of the electrical center may not be known or may be inaccessible, the connection is usually made at a convenient point at a distance X (Figure 2-19) from the electrical center, D. The distance from the true center of the electrode to the current probe (assuming the measurements are made on a radial from the electrical center) is  $C_k + X$ . The use of 62 percent point on the curves of Figure 2-18 to determine the resistance of the earth electrode should in reality correspond to a position of the potential probe that is 0.62 ( $C_k + X$ ) from the true center (D). This means that the distance,  $P_t$ , from the point of actual connection (O) to the system to the location at which the correct resistance to earth exists will be

$$\begin{aligned}
 P_t &= 0.62 (C_k + X) - X & (2-50) \\
 &= 0.62 C_k - 0.38 X
 \end{aligned}$$

where

$P_t$  = Distance of potential probe from point of connection to electrode when the measured resistance is the true value of resistance-to-earth for the electrode,

$C_k$  = Current probe distance from point of connection, for the kth set of probe measurements, and

$X$  = Distance from electrical center of electrode system to point of connection to the electrode system.

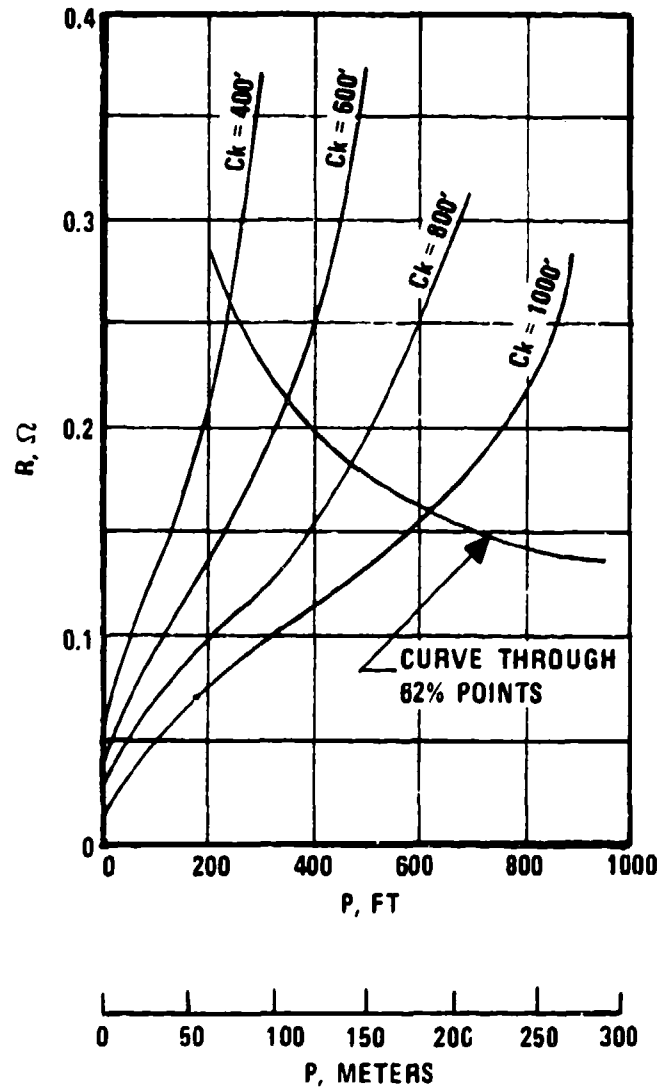


Figure 2-18. Earth Resistance Curves for a Large Electrode Subsystem

To determine the true resistance of the earth electrode,  $X$  is allowed to assume convenient increments from zero to  $C_K$ . For each  $C_K$ , the value of measured resistance corresponding to the resultant  $P_t$  (calculated with Equation 2-50) is read from the curves of Figure 2-18 and plotted against  $X$ . For example, if  $X$  and  $C_K$  both equal 305 m (1000 feet), considering only the right hand curve in Figure 2-18, the value of  $P_t$  is 240, and  $R$  is 0.08 ohms. Next let  $X$  be 244 m (800 feet). The corresponding value of  $P_t$  is 96 m (316 feet) and  $r$  is 0.1 ohms. In this manner, estimates of the 62 percent values can be taken from Figure 2-18 and replotted as "true" resistance versus  $X$ , as shown in Figure 2-20. At the region of intersection of the curves in Figure 2-20, the value of  $X = 122$  m (400 feet) corresponds to the electrical center of the electrode, and the corresponding value of resistance (0.13 ohms) is the true value of resistance-to-earth of the electrode system. It is recommended that the distance to the current probe, "C", from the point of connection to the earth electrode, "O", (see Figure 2-19) be between one and two times the length of the longest side of the electrode system. Furthermore, failure to obtain a well defined region of intersection of the curves can result if the probe measurements are not taken on a radial from the electrical center, in that case, new probe directions will be required.

**2.7.2.3 Test Equipments.** Test equipments are presently available which will permit the accurate measurement of ground resistances of earth electrode subsystems from 0.01 to 20,000 ohms and above. Most equipments used in conducting these measurements are designed to utilize ground test currents other than dc or 60 Hz to avoid or eliminate the effects of stray ac or dc currents in the earth.

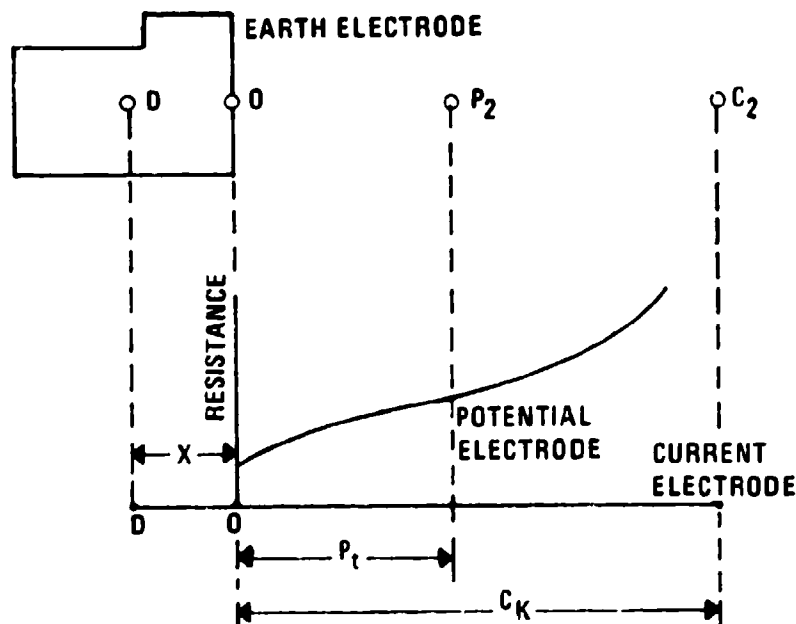


Figure 2-19. Earth Resistance Curve Applicable to Large Earth Electrode Subsystems

**2.7.3 Three-Point (Triangulation) Method.** In this method, illustrated in Figure 2-21 the resistances of the electrode under test ( $R_x$ ) and the auxiliary electrodes ( $R_a$ ,  $R_b$ ) are measured two at a time. The unknown resistance is then computed from the formula.

$$R_x = \frac{(R_x + R_a) + (R_x + R_b) - (R_a + R_b)}{2}, \quad (2-51)$$

where the terms in the parenthesis are the following measured resistances:

$$(R_x + R_a) = \frac{V_x^a}{I_x} \quad (2-52)$$

= voltage drop from test electrode, X, to electrode A, divided by current entering test electrode, X,

$$(R_x + R_b) = \frac{V_x^b}{I_x} \quad (2-53)$$

= voltage drop from test electrode to electrode B, divided by current into test electrode, X,

$$(R_a + R_b) = \frac{V_a^b}{I_a} \quad (2-54)$$

= voltage drop from electrode A to electrode B, divided by current entering electrode A.

For best accuracy, it is important to use auxiliary electrodes with resistances of the same order of magnitude as the unknown. The series resistances may be measured either with a bridge or with a voltmeter and ammeter. Either alternating or direct current may be used as the source of test current. For the three-point measurement, the electrodes must be at some distance from each other; otherwise absurdities such as zero or even negative resistances may arise in the calculations. In measuring a single 3 meter (10-foot) driven ground rod, the distance between the three separate ground electrodes should be at least 5 meters (15 feet), with a preferable spacing of 8 meters (25 feet) or more. For larger area grounds, which are presumable of lower resistances, spacing on the order of the dimensions of the ground field is required as a minimum. This method is most effective for measurement of single rods and is not recommended for multi-rod earth electrode subsystems.

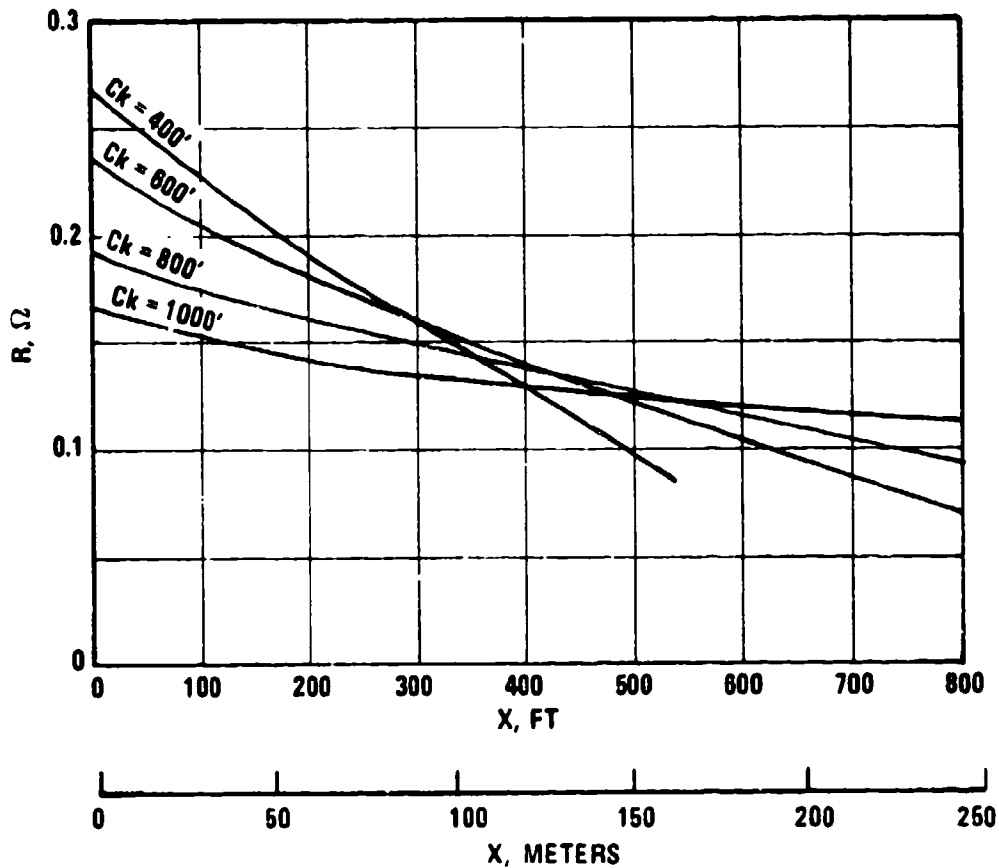


Figure 2-20. Intersection Curves for Figure 2-18

## 2.8 OTHER CONSIDERATIONS.

**2.8.1 Surface Voltages Above Earth Electrodes.** Very large currents can be conducted into earth electrodes whenever power line faults or lightning strikes occur. As a result, there is a substantial voltage developed at the surface of the earth near the electrode; this voltage varies significantly with distance from the electrode connection point. The voltage difference between two points about three feet apart on the surface is the "step voltage", i.e., it is the voltage level between the feet of a person standing or walking on the surface.

**2.8.1.1 Step Voltage Safety Limit.** The maximum safe step voltage depends upon the duration of the individual's exposure to the voltage and upon the resistivity of the earth at the surface. The maximum safe step voltage for a shock duration of from 0.03 to 3.0 seconds has been expressed (2-3) as

$$V_{\text{step (safe)}} = \frac{165 + \rho_s}{\sqrt{t}} \quad (2-55)$$

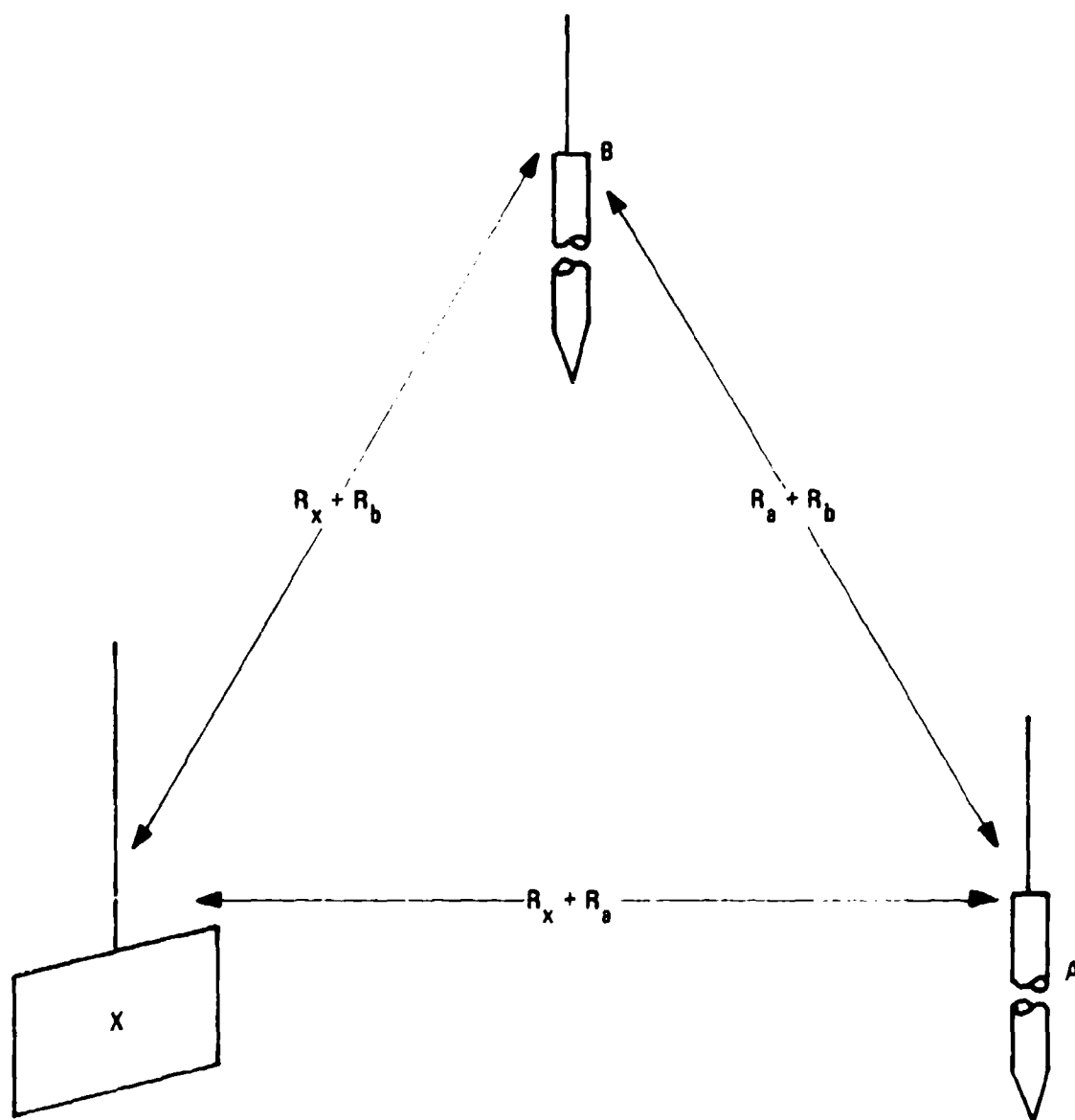


Figure 2-21. Triangulation Method of Measuring the Resistance of an Earth Electrode

where

$\rho_s$  = surface earth resistivity, (ohm - meters),

$\approx 10$  for a minimum value,

$t$  = duration of shock (sec).

For a 30 millisecond or shorter duration, the maximum safe step voltage is 1000 volts, and for durations greater than 3 seconds, it is 100 volts.

**2.8.1.2 Step Voltages for Practical Electrodes.** The expressions for step voltage estimates in homogeneous soil for both flush and buried vertical rod electrodes and for buried grid electrodes are given in the following paragraphs. It should be noted that step voltages depend upon electrode geometry as well as upon earth resistivity and current magnitude.

**2.8.1.2.1 Flush Vertical Rod.** The potential on the earth at a distance  $x$  from the top of a single, isolated flush-driven vertical rod is (2-3)

$$V_x = \frac{0.366 \rho I_o}{\ell} \log \left( \frac{\ell}{x} + \sqrt{1 + \frac{\ell^2}{x^2}} \right), \quad (2-56)$$

and the potential of the rod itself is

$$V_o = \frac{0.366 \rho I_o}{\ell} \log \frac{3\ell}{d}. \quad (2-57)$$

The step potential at the ground rod (where  $p$  is equal to a pace, or step, length from the rod) is therefore

$$\begin{aligned} V_o - V_p &= \frac{0.366 \rho I_o}{\ell} \left[ \log \frac{3\ell}{d} - \log \left( \frac{\ell}{p} + \sqrt{1 + \frac{\ell^2}{p^2}} \right) \right], \quad (2-58) \\ &= \frac{0.366 \rho I_o}{\ell} \log \left[ \frac{3p}{d \left( 1 + \sqrt{\frac{p^2}{\ell^2} + 1} \right)} \right]. \end{aligned}$$

When the step length is much less than the rod length, i.e., when  $p \ll \ell$ , the step voltage can be approximated as

$$V_o - V_p \approx \frac{0.366 \rho I_o}{\ell} \log \left( \frac{3p}{2d} \right) \quad (2-59)$$

The step potential can be expressed as a fraction of the ground rod potential as follows:

$$\frac{V_o - V_p}{V_o} = \frac{\log \left[ d \left( 1 + \sqrt{\frac{3p^2}{\ell^2} + 1} \right) \right]}{\log \frac{3\ell}{d}} \quad (2-60)$$

The fractional step voltages for ground rods of various length are given in Table 2-7. For this Table, rod diameter is assumed to be one inch (2.54 cm) and the pace length is assumed to be three feet (0.91 m).

Table 2-7

## Step Voltages for a Buried Vertical Ground Rod

<u>Rod Length</u> (Ft)	<u>Ratio of Step Voltage</u> <u>To Electrode Potential</u>
5	0.75
10	0.68
20	0.61
50	0.53
100	0.48



The step voltage near the 10-foot by 1-inch (3.05 m x 2.54 cm) rod in 1000 ohm-cm soil is about 68 percent of the voltage between the rod and a point approaching an infinite distance away. Step voltage near a ground rod will be between 80 percent of the rod potential (for very short rods) and 50 percent (for very long rods).

The step voltage on the surface of the earth near an isolated 10-foot by 1-inch (3.05 m x 2.54 cm) ground rod ( $\rho = 1000$  ohm-cm) carrying a lightning current of 20,000 amps could be fatal since the step voltage would be

$$V_o - V_p = \frac{(0.366)}{(10)} \frac{(1000)}{(12)} \frac{(20,000)}{(2.54)} \quad (2-81)$$

$$\begin{aligned} & \times \left[ \log 360 - \log \left( \frac{10}{3} + \sqrt{1 + \frac{100}{9}} \right) \right], \\ & = \frac{7.32 \times 10^6}{3.05 \times 10^2} [2.556 - 0.833] \\ & = (2.4 \times 10^4) (1.723) \\ & = 41,352 \text{ volts,} \end{aligned}$$

which is 41 times higher than the safe step voltage derived above.

The resistance of the 10-foot by 1-inch (3.05 m x 2.54 cm) rod in 1000 ohm-cm soil is

$$\begin{aligned} R &= \frac{0.366 \rho}{L} \log \frac{3L}{d}, \quad (2-82) \\ &= \frac{(0.366)}{(10)} \frac{(10^3)}{(12)} \frac{(2.54)}{(2.54)} \log 360, \\ &= 1.2 \log 360, \\ &= (1.2) (2.556), \\ &= 3.1 \text{ ohms.} \end{aligned}$$

Higher values of earth resistivity would cause the step voltage near the rod to be even higher than the calculated 41,400 volts. For a three second duration shock condition, the requirement that the step voltage not exceed 100 volts means that the single 10-foot by 1-inch (3.05 m x 2.54 cm) rod would produce an unsafe step voltage with a fault current greater than about 50 amperes, even in low resistivity (1000 ohm-cm) soil.

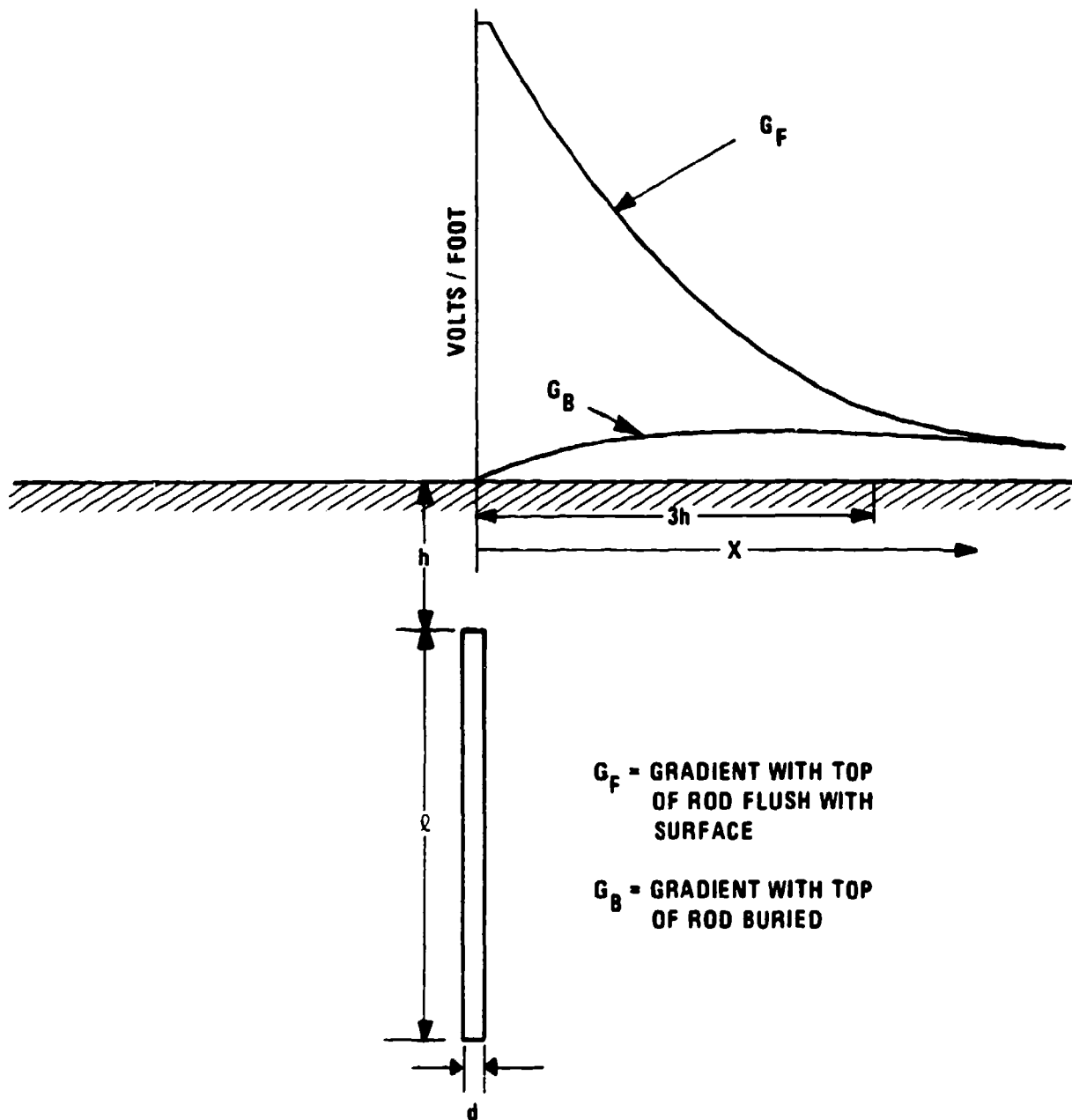


Figure 2-22. Variation of Surface Potential Produced by a Current Flowing Into an Isolated Ground Rod

**2.8.1.2.2 Buried Vertical Rod.** If the single isolated vertical rod is driven so that the top of the rod is below the surface, the maximum step voltage on the surface of the earth is reduced. Figure 2-22 shows the surface voltage variation for a flush driven rod compared with that for a rod with its top below the surface. Maximum gradient for the flush driven rod is at the vicinity of the rod. Maximum gradient for the rod sunk into the earth to a depth of  $h$  feet occurs at a distance of  $3h$  to  $4h$  from the rod (2-3). The step voltage for the rod driven so that its top is  $h$  feet below the surface is:

$$V_x - V_{x+p} = \frac{0.366\rho I_o}{\ell} \log \left\{ \frac{\left[ \sqrt{(h+\ell)^2 + x^2} + h + \ell \right] \left[ \sqrt{h^2 + (x+p)^2} + h \right]}{\left[ \sqrt{(h+\ell)^2 + (x+p)^2} + h + \ell \right] \left[ \sqrt{h^2 + x^2} + h \right]} \right\} \quad (2-63)$$

For

$$\rho = 10^3 \text{ ohm-cm,}$$

$$I_o = 20,000 \text{ amperes,}$$

$$\ell = 10 \text{ feet (3.05 m),}$$

$$x = 3h \text{ feet, and}$$

$$p = 3 \text{ feet (0.91 m)}$$

$$V_x - V_{x+p} = 2.4 \times 10^4 \log \left\{ \frac{\left[ \sqrt{(h+10)^2 + (3h)^2} + h + 10 \right] \left[ \sqrt{h^2 + (3h+3)^2} + h \right]}{\left[ \sqrt{(h+10)^2 + (3h+3)^2} + h + 10 \right] \left[ \sqrt{h^2 + (3h)^2} + h \right]} \right\} \quad (2-64)$$

If  $h = 3$  feet, the maximum step voltage is approximately

$$\begin{aligned} V_x - V_{x+p} &= 2.4 \times 10^4 \log \left\{ \frac{(28.81)(15.37)}{(30.69)(12.49)} \right\} \\ &= 2.4 \times 10^4 \log 1.15 \\ &= (2.4 \times 10^4) 0.063 \\ &= 1504 \text{ volts,} \end{aligned}$$

instead of 41,400 volts, which was characteristic of the flush-driven rod.

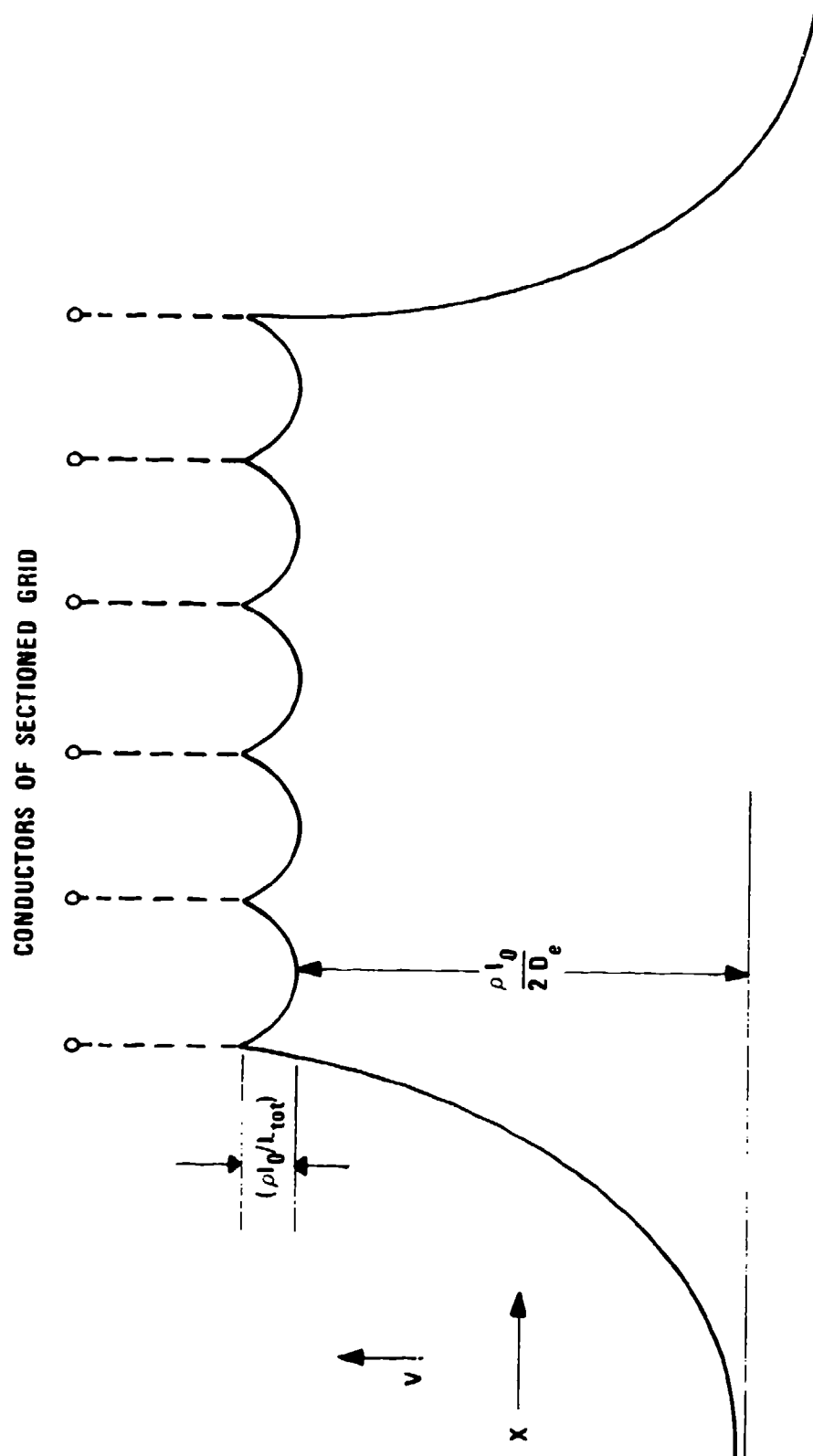


Figure 2-23. Surface Potential Variation Along a Grid

2.8.1.2.3 Buried Horizontal Grid. An expression for the resistance to earth for a buried grid was presented in Section 2.6.2.3. Equations 2-27 and 2-28 are the sum of a resistance of a superficial plate ( $\rho/2D_e$ ) and a resistance term representing the per unit diffusion resistance of the earth electrode material ( $\rho/L$ ). A voltage  $\rho I/L$  which is proportional to the per unit average current flowing from the conductors of the mesh into the earth represents an approximation of the potential difference between the conductors of the mesh and the center of the open space with each mesh. The sketch of Figure 2-23 shows the resultant voltage distribution across a section of a grid. Note that the approximation used here would predict that

$$\frac{\rho I_0}{2D_e} \quad (2-65)$$

is the minimum voltage (with respect to infinity) at the edge of the grid, so that the grid simply translates the dangerous voltage gradient to the periphery of the grid (2-3).

If the value of earth resistivity is moderately high--say  $10^4$  ohm-cm--and if the lightning current is  $2 \times 10^4$  amperes, the grid in the example of Section 2.6.2.4 would exhibit

$$\begin{aligned} \frac{\rho I}{L} &= \frac{(10^4)(2 \times 10^4)}{6.7 \times 10^4} \\ &= 3000 \text{ volts} \end{aligned} \quad (2-66)$$

over a five-foot (1.5 m) distance. This would exceed the safe step voltage of 1000 volts, developed earlier.

If the grid is made of conductors spaced one foot apart for a total conductor length of 20,200 feet (6157 m) there would be 10,000 meshes on the 10,000 square foot (929 m<sup>2</sup>) area. The effective diameter would still be 113 feet (34.4 m), and the computed resistance would be

$$\begin{aligned} R_e &= \rho \left[ \frac{1}{2(3440)} + \frac{1}{(2.02 \times 10^4)(12)(2.54)} \right] \\ &\approx 1.45 \times 10^{-4} \rho + \frac{10^{-4}}{61} \rho \end{aligned} \quad (2-67)$$

The maximum step potential difference over the grid of the latter case, again assuming  $\rho$  is  $10^4$  ohm-cm and an effective lightning current of 20,000 amperes, would be

$$\frac{\rho I}{L} = \frac{(10^4)(2 \times 10^4)}{(62)(10^4)}$$

$$= 322 \text{ volts}$$

This would be a safe value of step voltage for transients shorter than 30 milliseconds, if the transient, or surge, impedance of the line does not greatly exceed its steady state resistance.

2.8.1.3 Minimizing Step Voltage. Table 2-8 lists several design approaches to reducing the potential hazards of step voltage. The most effective method is the reduction of the resistance to earth of the earth electrode system to as low a value as is economically feasible.

Table 2-8

## Methods of Reducing Step Voltage Hazards

<u>Design Approach</u>	<u>Remarks</u>
1. Minimize resistance to earth of electrode system.	Resistance to earth is directly proportional to soil resistivity.
2. Bury earth electrode to reduce maximum gradient on surface of earth.	Connection to earth electrode must be insulated to withstand $5 \times 10^4 R_0$ volts.
3. Bury a grid beneath the earth, surrounding the earth electrode.	Tends to equalize the surface potential over area of grid.
4. Erect barricade so that personnel cannot enter area of danger.	Fence must be grounded.

**2.8.2 Heating of Electrodes.** It is necessary to use enough material in an earth electrode to prevent excessive local heating when large currents flow in the electrode.

**2.8.2.1 Steady State Current.** The presence of fault current in the earth electrode subsystem must be limited to a value which will not raise the temperature of the soil above the boiling point of water. The tolerable steady state ac current into an earth electrode is (2-3).

$$I_{ss} = \frac{1}{R} \sqrt{0.024 \rho \Delta T} , \quad (2-68)$$

where

$\rho$  = earth resistivity,  
 $R$  = electrode resistance to earth, and  
 $\Delta T$  = permissible temperature rise ( $^{\circ}\text{C}$ ).

For  $\Delta T = 60^{\circ}\text{C}$ , the permissible steady state current is limited by

$$I_{ss} < \frac{1200}{R} \text{ amp}$$

when  $\rho = 10^6 \text{ ohm-cm}$ , and by

$$I_{ss} < \frac{38}{R} \text{ amp}$$

when  $\rho = 10^3 \text{ ohm-cm}$ . Since the voltage at the earth electrode is equal to the product  $I R$ , the corresponding voltage limits are

$$E_{ss} < 1200 \text{ volts, } \rho = 10^6 \text{ ohm-cm}$$

and

$$E_{ss} < 38 \text{ volts, } \rho = 10^3 \text{ ohm-cm.}$$

**2.8.2.2 Transient Current.** The permissible transient current density for a temperature rise that does not exceed  $60^{\circ}\text{C}$  is found from the transient temperature time expression (2-3)

$$i = \sqrt{\frac{1.75 \Delta T}{\rho t}} \text{ amp/cm}^2 , \quad (2-69)$$

where

$t$  = duration of the transient, in seconds,

$\Delta T$  = temperature rise,

$\rho$  = soil resistivity, ohm-cm, and

$i$  = transient current density.

Letting  $\Delta T = 60^\circ\text{C}$ , one has

$$i_{\max} = 10 \sqrt{\frac{1}{\rho t}} \quad \text{amp/cm}^2$$

The current density,  $i$ , at the surface of a short ground rod is approximately constant over the length of the rod and is given by

$$i = \frac{I}{\pi d l} \quad \text{amp/cm}^2, \quad (2-70)$$

where

$d$  = rod diameter (cm),

$l$  = rod length (cm), and

$I$  = input current (amperes).

For a 10-foot by 1-inch rod (3.05 m x 2.54 cm), the peak transient current which can be handled without causing greater than  $60^\circ\text{C}$  temperature rise is:

$$\begin{aligned} I_{\max} &= \pi d l i_{\max} \\ &= \pi (2.54) (304.8) (10) \sqrt{\frac{1}{\rho t}} \\ &= 2.43 \times 10^4 \sqrt{\frac{1}{\rho t}} \quad \text{amperes.} \end{aligned} \quad (2-71)$$



**2.8.2.3 Minimum Electrode Size.** The necessity to hold the surface temperature below boiling temperatures establishes a minimum amount of electrode material.

The minimum length of a single ground rod is

$$l \geq \frac{I \sqrt{\rho t}}{10 \pi d} \text{ cm.} \quad (2-72)$$

The value of  $I \sqrt{t}$  is approximately 1000 for both lightning stroke currents and power system fault currents, so for satisfactory energy dissipation the minimum rod length is specified by

$$l \geq \frac{32}{d} \sqrt{\rho} \text{ cm.} \quad (2-73)$$

If the earth is moist soil with a  $\rho$  of approximately  $10^4$  ohm-cm, the limit becomes

$$l \geq \frac{3200}{d} \text{ cm.} \quad (2-74)$$

In granite with a  $\rho$  of approximately  $10^6$  ohm-cm, the limit becomes

$$l \geq \frac{32000}{d} \text{ cm.} \quad (2-75)$$

If 2-cm rods are used, the safe dissipation of heat in granite would require at least 80 rods, each 2 meters long. For moist earth, only 8 rods, each 2 meters long, would be required for heat dissipation.

## 2.9 ELECTRODE ENHANCEMENT.

**2.9.1 Introduction.** Sites may be encountered where acceptable and practical numbers of driven rods, buried cables, and other available materials will not achieve the desired low resistance to earth for special communication systems, i.e., HF transmitters. In such situations, enhancement of the resistivity of the soil around the electrodes may be necessary to lower the resistance to the desired value. While enhancement of the resistivity may be required in certain situations, discretion of its use should be exercised due to the reduced life span of the earth electrode subsystem.

The resistance to earth of an electrode is directly proportional to soil resistivity and inversely proportional to the total area of contact established with the soil. For fixed land areas, additional vertical rods or horizontal cables produce diminishing returns because of increased mutual coupling effects. The most straight forward enhancement method is to reduce soil resistivity. The parameters which strongly affect soil resistivity are moisture content, ionizable salt content, and porosity; the latter determining the moisture retention properties of the soil. Thus two recommended techniques for reducing earth resistivity are water retention and chemical salting.

**2.9.2 Water Retention.** Overdrainage of soil leaches away salts that are necessary for high conductivity and dries out the deeper layers, thereby increasing their resistivity. Planting of appropriate ground covers, such as legumes, to retard runoff and to enhance the natural production of salts in the soil is useful. Surface drainage should be channeled so as to keep the earth electrode subsystem moist. Maintaining moist earth over the extent of the earth electrode subsystem will keep soil salt in solution as conductive ions. Drainage water which is high in salt content can be useful for continuous salting of the earth electrode.

A porous clay, bentonite (also known as well drillers mud) can absorb water from surrounding soil and has hydration as well as water retention properties. When placed around ground rods and their interconnecting cable, it greatly increases the effective area of the rod and cable which in turn reduces the resistance of the earth electrode subsystem to earth (2-14, 2-15). Bentonite is generally available in dry (powder) form, must be saturated with water after initial installation and should be topped with a 12-inch layer of excavated soil. Caution is urged when using bentonite in areas that will ultimately be paved as it can expand to several times its dry volume when saturated. This can also prove to be a disadvantage of bentonite since it expands and contracts so much with moisture content, it can pull away from the ground rod and surrounding soil when moisture is lost. A much better backfill around ground rods is a mixture of 75 percent gypsum, 20 percent bentonite clay, and 5 percent sodium sulfate. The gypsum, which is calcium sulfate, absorbs and retains moisture and adds reactivity and conductivity to the mixture. Since it contracts very little when moisture is lost, it will not pull away from the ground rod or surrounding earth. The bentonite insures good contact between ground rod and earth by its expansion, while the sodium sulfate prevents polarization of the rod by removing the gases formed by current entering the earth through the rod. This mixture is available from cathodic protection distributors as standard galvanic anode backfill and is relatively inexpensive. The backfill mixture should be covered with 12 inches of excavated soil. This mixture is superior to chemical salts since it is much more enduring.

**2.9.3 Chemical Salting.** Reduction of the resistance of an electrode may also be accomplished by the addition of ion-producing chemicals to the soil immediately surrounding the electrode. The better known chemicals in the order of preference are:

- a. Magnesium sulphate ( $MgSO_4$ ) - epsom salts.
- b. Copper sulphate ( $CuSO_4$ ) - blue vitriol.
- c. Calcium chloride ( $CaCl_2$ ).
- d. Sodium chloride ( $NaCl$ ) - common salt.
- e. Potassium nitrate ( $KNO_3$ ) - saltpeter.

Magnesium sulphate (epsom salts), which is the most common material used, combines low cost with high electrical conductivity and low corrosive effects on a ground electrode or plate. The use of common salt or saltpeter is not recommended as either will require that greater care be given to the protection against corrosion. Additionally, metal objects nearby but not related to grounding will also have to be treated to prevent damage by corrosion. Therefore, salt or saltpeter should only be used where absolutely necessary.

Large reductions in the resistance to earth of the individual ground electrodes may be expected after chemical treatment has been applied to the earth. The initial effectiveness of chemical treatment is greatest where the soil is somewhat porous because the solution permeates a considerable volume of earth and increases the effective size of the electrode. In compact soils, the chemical treatment is not as immediately effective because the material tends to remain in its original location for a longer period of time.

The effectiveness of chemical treatment in lowering the resistance of a ground rod is illustrated by Figures 2-24 and 2-25. Chemical treatment achieves a significant initial reduction of resistance and further stabilizes the resistance variations. It also limits the seasonal variation of resistance and, additionally, lowers the freezing point of the surrounding soil.

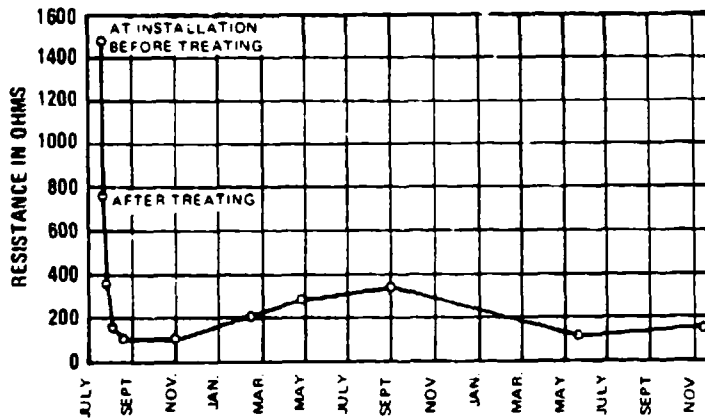


Figure 2-24. Effect of Chemical Treatment on Resistance of Ground Rods

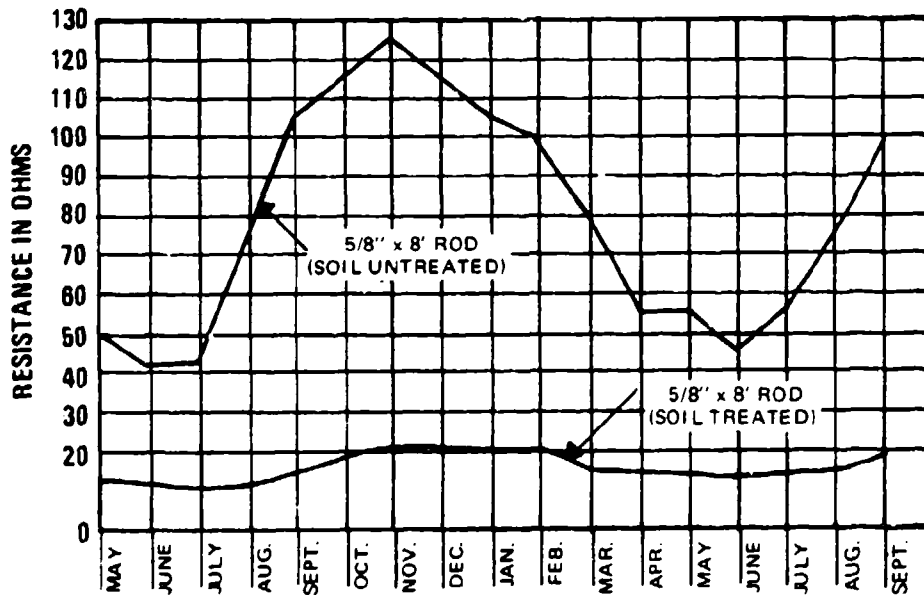


Figure 2-25. Seasonal Resistance Variations of Treated and Untreated Ground Rods

Chemical treatment is limited in its effectiveness, however. Consider, for example, a square array of 100 ground rods of length  $l$  with spacings of twice the length of a rod. The resistance to earth (using an extrapolated value of 3 for  $K$ ) from Figure 2-12 is (see also Equations 2-16 and 2-26)

$$R = \frac{(3) (0.366) \rho}{100 l} \log \frac{3l}{d}$$

Assuming that

$$\rho = 10^6 \text{ ohm-cm (gravel sand stone),}$$

$$l = 100 \text{ feet (30.5 m) per rod, and}$$

$$d = 1 \text{ inch (2.54 cm),}$$

then

$$R = \frac{(3) (0.366) (10^6) (3.56)}{(100) (100) (12) (2.54)}$$

$$\approx 12.81 \text{ ohms}$$

The upper bound on the effectiveness of chemical enhancement can be illustrated by determining the resistance to earth of a metal electrode which would completely fill the volume of earth (1800 x 1800 x 100 ft., i.e., 550 x 550 x 30 m) occupied by the above array of ground rods. The effective diameter,  $D_e$ , of the equivalent plate would be 2030 feet (619 m), and its resistance to earth would be (2-3):

$$R \approx \frac{\rho}{2D_e}$$

$$= \frac{10^6}{(2) (2030) (12) (2.54)} \quad (2-76)$$

$$\approx 8 \text{ ohms}$$

The most that chemical enhancement could reduce the resistance of this large array would be by a factor of 1.58.

**2.9.4 Electrode Encasement.** The calculations of resistance of earth electrodes invariably assume zero contact resistance between the electrode elements and the earth. In reality, however, the interface between the surface of the rod and the earth is far from uniform except when the earth is tamped clay or its equivalent. Granular earth (gravel, etc.) makes very poor contact. Reduction of this contact resistance should have a strong effect on reducing the electrode resistance because it is close to the electrode where current density is high. Encasing the electrode in conductive mastic or conductive concrete is one approach to improving the contact between the electrode and the earth. Effects of local variations or moisture content will also be reduced and stabilized, if the encasement material absorbs and holds moisture.

**2.9.5 Salting Methods.** The trench method for treating the earth around a driven electrode is illustrated in Figure 2-26. A circular trench is dug about one foot deep around the electrode. This trench is filled with the soil treating material and then covered with earth. The material should not actually touch the rod in order to provide the best distribution of the treating material with the least corrosive effect.

Another method for treating the earth around a driven electrode, using magnesium sulphate and water, is illustrated in Figure 2-27. A 2-foot length (approximately) of 8-inch diameter tile pipe is buried in the ground surrounding the ground electrode. This pipe is then filled with magnesium sulphate to within one foot of grade level and watered thoroughly. The 8-inch tile pipe should have a wooden cover with holes and be located at ground level.

None of the aforementioned chemical treatments permanently improve earth electrode resistance. The chemicals are gradually washed away by rainfall and through natural drainage. Depending upon the porosity of the soil and the amount of rainfall, the period for replacement varies. Forty to ninety pounds of chemical will initially be required to maintain effectiveness for two or three years. Each replenishment of chemical will extend the effectiveness for a longer period so that the future treatments have to be done less and less frequently.

Another method of soil treatment or electrode enhancement involves the use of hollow made electrodes which are filled with materials/salts which absorb external atmospheric moisture. These electrodes (generally 8-feet long) must be placed in holes drilled by an earth auger making sure the breather holes at the top are above grade level. Moisture from the atmosphere is converted to an electrolyte which in turn seeps through holes in the electrode into the surrounding soil. This keeps the soil moist and thereby reduces the resistance of the electrode to earth. These electrodes should be checked annually to ensure sufficient quantities of materials/salts are available and that good continuity exists between the rod and interconnecting cable.

## **2.10 CATHODIC PROTECTION.**

**2.10.1 Introduction.** When two metals of different types are immersed in wet or damp soil, a basic electrolytic cell is formed. A voltage equal to the difference of the oxidation potentials of the metals will be developed between the two electrodes of the cell. If these electrodes are connected together through a low resistance path, current will flow through the electrolyte with resultant erosion of the anodic member of the pair. Unfortunately, those factors that aid in the establishment of low resistance to earth also foster corrosion. Low resistance soils with a high moisture level and a high mineral salt content provide an efficient electrolytic cell with low internal resistance. Relatively large currents can flow between short-circuited electrodes (such as copper ground rods connected to steel footings or reinforcing rods in buildings) and quickly erode away the more active metal (see Section 7.8.1.2) of the cell. In high-resistance cells, the current flow is less and the erosion is less severe than in low-resistance cells.

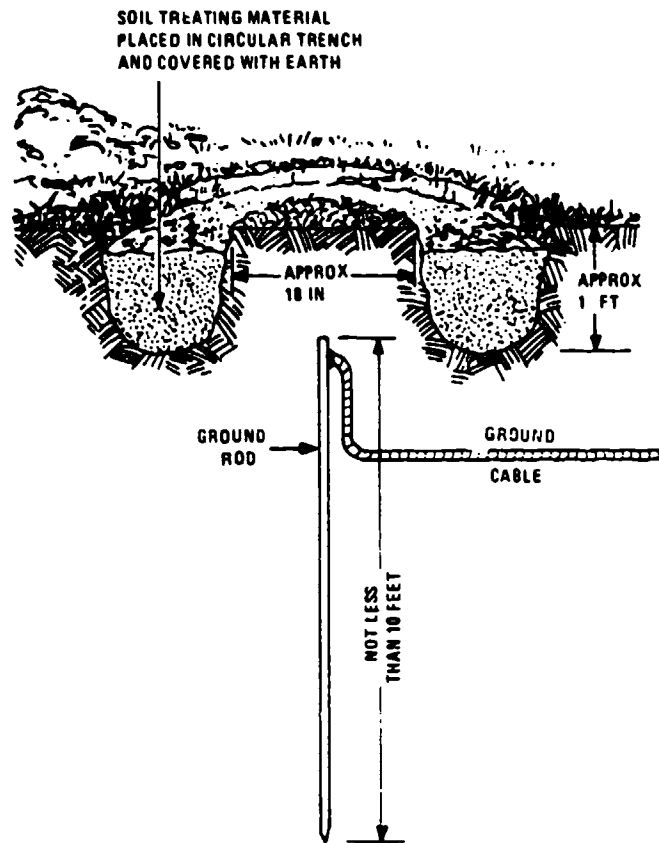


Figure 2-26. Trench Method of Soil Treatment

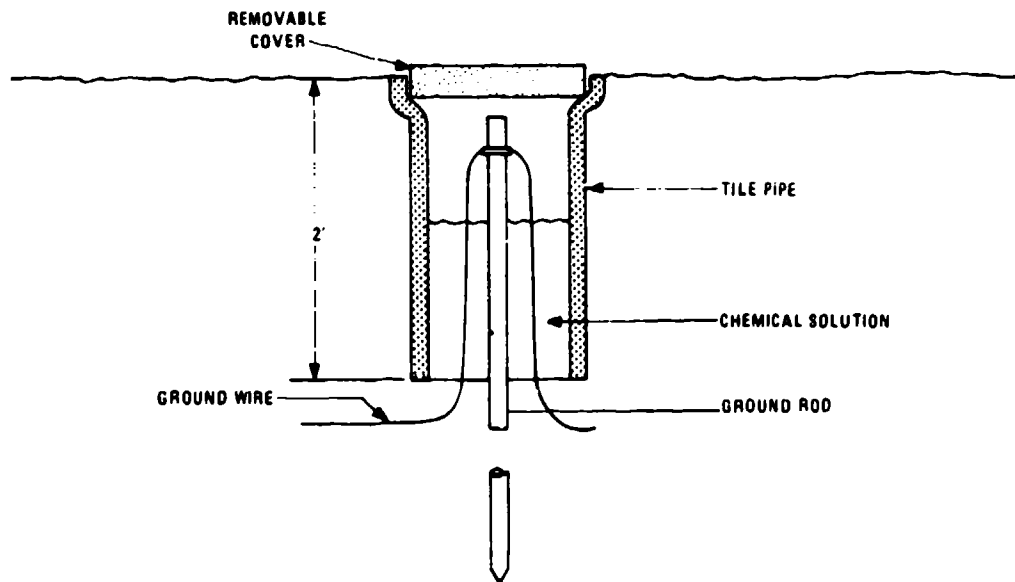


Figure 2-27. Alternate Method of Chemical Treatment of Ground Rod

### 2.10.2 Protection Techniques.

Three basic techniques can be used to lessen the corrosion rate of buried metals. The obvious method is to insulate the metals from the soil by the use of protective coatings. This interrupts the current path through the electrolyte and stops the erosion of the anode. Insulation, however, is not an acceptable corrosion preventive for earth electrodes. The second technique for reducing galvanic corrosion is avoiding the use of dissimilar metals at a site. For example, if all metals in contact with the soil are of one type (such as iron, lead or copper), galvanic corrosion is minimized. Each of these materials, however, has unique properties such as weight, cost, conductivity, ductility, strength, etc., that makes its use desirable, and thus none can be summarily dismissed from consideration for underground applications. Copper is a desirable material for the earth electrode subsystem; apart from its high conductivity, the oxidation potential of copper is such that it is relatively corrosion resistant. Since copper is cathodic relative to the more common structural metals, its corrosion resistance is at the expense of other metals. Iron electrodes would, of course, be compatible with water pipes, sewer lines, reinforcing rods, steel pilings, manhole covers, etc., but iron is subject to corrosion even in the absence of other metals. In addition, the conductivity of iron is less; however, steel grounding rods are sometimes used by electric utilities for grounding associated with their transmission lines. Because of the greater conductivity and corrosion resistance of copper, it is normally used for the grounding of buildings, substations, and other facilities where large fault or lightning currents may occur and where voltage gradients must be minimized to ensure personnel protection.

The third technique for combating the corrosion caused by stray direct currents and dissimilar-metal unions is commonly called cathodic protection. Cathodic protection may be implemented through the use of sacrificial anodes or the use of an external current supply to counteract the voltage developed by oxidation. Sacrificial anodes containing magnesium, aluminum, manganese, or other highly active metal can be buried in the earth nearby and connected to an iron piling, steel conduit, or lead cable shield. The active anodes will oxidize more readily than the iron or lead and will supply the ions required for current flow. The iron and lead are cathodic relative to the sacrificial anodes and thus current is supplied to counteract the corrosion of the iron or lead. The dc current is normally derived from rectified alternating current, but occasionally from photovoltaic cells, storage batteries, thermoelectric generators, or other dc sources. Since the output voltage is adjustable, any metal can be used as the anode, but graphite and high silicon iron are most often used because of their low corrosion rate and economical cost. Cathodic protection is effective on either bare or coated structures. If the sacrificial anodes are replenished at appropriate intervals, the life of the protected elements is significantly prolonged.

**2.10.3 Sacrificial Anodes.** Sacrificial anodes provide protection over limited areas. Impressed current cathodic protection systems use long lasting anodes of graphite, high silicon cast iron or, to a lesser extent, platinum coated niobium or titanium. The protection of long cable or conduit runs can be provided by biasing the metal to approximately -0.7 to -1.2 volts relative to the surrounding soil. The external dc source supplies the ionization current that would normally be provided by the oxidation of the cable sheath or conduit. This dc current is normally derived from rectified ac and occasionally from photovoltaic cells, storage batteries, thermoelectric generators, or other dc sources. A layer of insulation such as neoprene must cover the metal to prevent direct contact with the surrounding soil. Therefore, the technique is not appropriate for protecting foundations, manholes, or other structural elements normally in contact with the soil. It is most appropriate for supplying the leakage current that would normally enter the soil through breaks in the insulation caused by careless installation, settling, lightning perforation, etc.

2.10.4 Corrosive Atmospheres. In regions exhibiting low soil resistivity, in corrosive atmospheres such as might be encountered near seashores, or near sources of large direct currents such as electroplating facilities, cathodic protection may be necessary to prolong the life of foundations, underground cable facilities, or other elements of a facility in contact with the soil. For additional information on the galvanic series of common metals see Table 7-7.

## 2.11 GROUNDING IN ARCTIC REGIONS.

2.11.1 Soil Resistivity. The problem of electric earth grounding in cold regions is primarily one of making good contact with high resistivity soils. Where frozen high resistivity materials are encountered, optimum grounding of power and communication circuits can only be accomplished by special attention to both surface and subsurface terrain. The fact is that resistance of frozen soils can be ten to a hundred times higher than in the unfrozen state. Seasonal changes in temperature and moisture will therefore extensively affect the soil resistance. Table 2-9 provides information on the effect of moisture content on earth resistivity, while Table 2-10 provides the effect of temperature on earth resistivity (2-16, 2-17).

Table 2-9. Effect of Moisture Content on Earth Resistivity

Moisture Content % By Weight	Resistivity, ohm-cm	
	Top Soil	Sandy Loam
0	1,000 x 10 <sup>4</sup>	1,000 x 10 <sup>6</sup>
2.5	250,000	150,000
5	165,000	43,000
10	53,000	18,500
15	17,000	10,500
20	12,000	6,300
30	6,400	4,200

Table 2-10. Effect of Temperature on Earth Resistivity \*

Temperature		Resistivity ohm-cm
°C	°F	
20	68	7,200
10	50	9,900
0	32 (water)	13,800
0	32 (ice)	30,000
-5	23	79,000
-15	5	330,000

\*For sandy loam, 15.2% moisture.



Permafrost occurs in various degrees throughout much of the arctic and subarctic regions and is defined as that part of the lithosphere (upper crust of the earth) in which a naturally occurring temperature below 0°C (32°F) has existed continuously for two or more years. The "annual frost zone" is the zone of annual freezing and thawing. Where permafrost occurs, the thickness of this surface layer varies from less than a foot in the arctic to depths in excess of 12 feet in the subarctic. The seasonal thaw zone remains unfrozen only during the short summer months. During this period, it is possible to recognize terrain features which can be located in the spring and fall if there is little or no snow cover.

Willow groves or aspen generally point to the absence of permafrost and to the presence of groundwater which freezes only for a short time. River bottoms and lake bottoms are usually frost-free. Generally, slow moving rivers and streams freeze from the top down (surface ice). Clear, fast moving rivers and streams usually freeze from the bottom up (anchor ice). Mountains, valleys, lake bottoms, streambeds, tree-covered slopes, tundra plains, swamplands, ice glaciers, silty estuaries, permafrost areas, and seasonably frozen ground, each will be found to affect soil resistivity. Consequently, it is easily seen how one area versus another might be more suitable for good grounding. Basic illustrations of variations, layering and asymmetrical contouring can be found in Figures 2-28, 2-29, and 2-30.

Resistance to ground and configuration of electrodes are further parameters that must be considered. The conductivity of cables and overhead wire systems are relatively high in comparison to the earth. Without the presence of minerals, dissolved salts, and moisture, clean dry soil can be classified as an insulator and possesses the intermediate characteristics of a poor conductor.

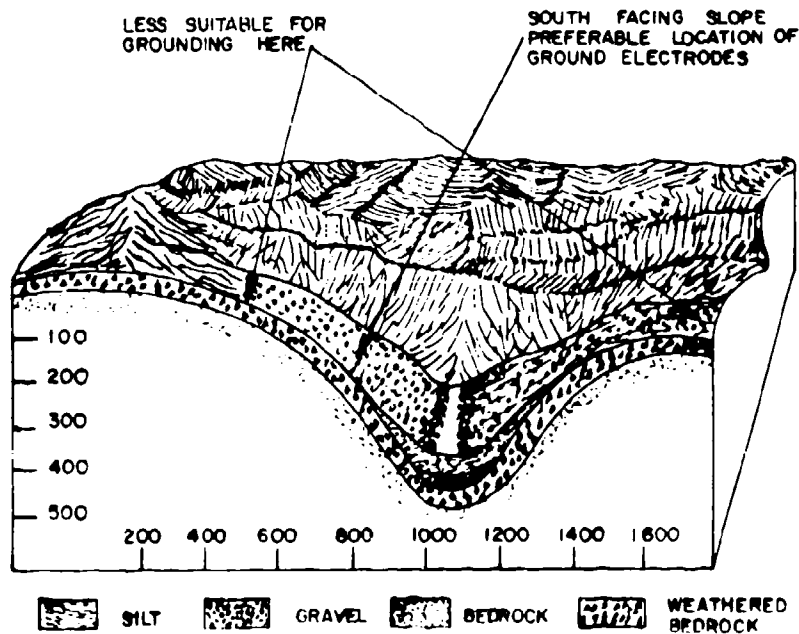


Figure 2-28. Relative Depths of Unconsolidated Materials, Subarctic Alaska

Seasonal freezing accounts for a reduced conductivity as illustrated in Table 2-10. If frozen soil or earth has a low conductivity, then providing larger effective electrodes will reduce the ground resistance. In northern arctic areas generally having very shallow surface thaw layers, horizontal rods or wires might be easier to install than driven rods and still provide optimum resistance values to earth or ground. Whether to install multiple electrodes or single deep, driven rods, or horizontal wires, the decision will usually be dependent on soil types and the economics of placement.

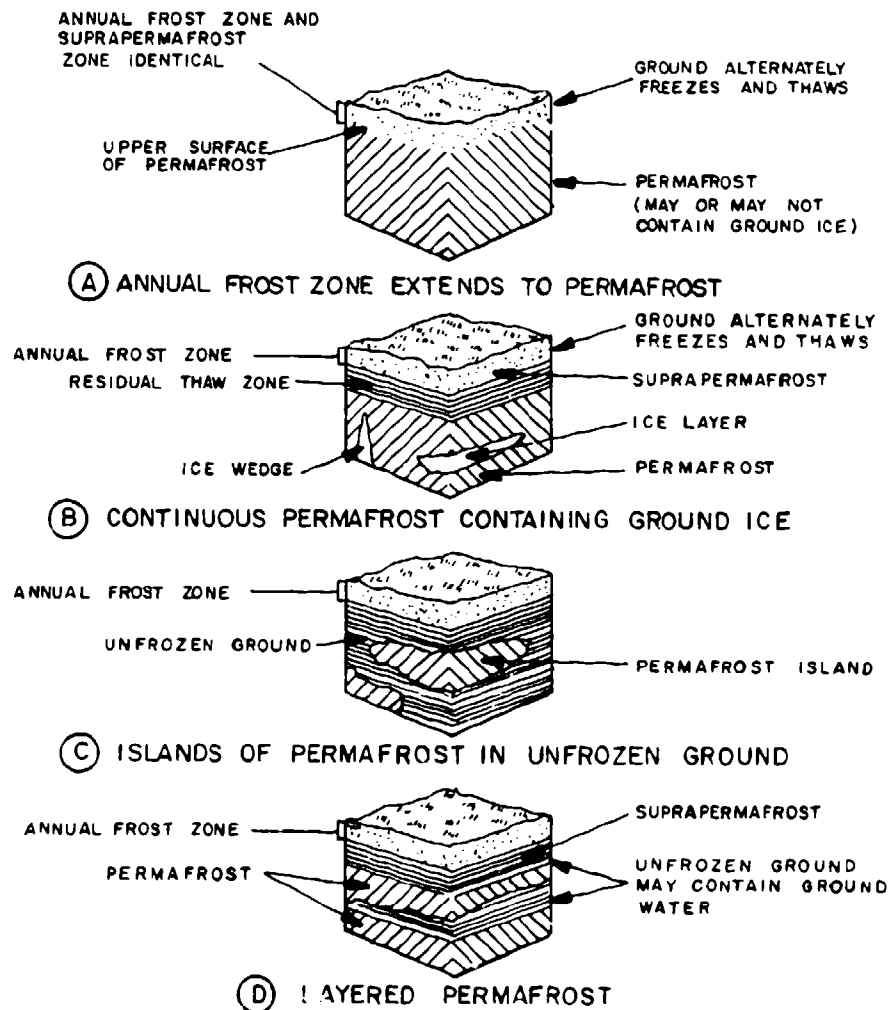


Figure 2-29. Typical Sections Through Ground Containing Permafrost

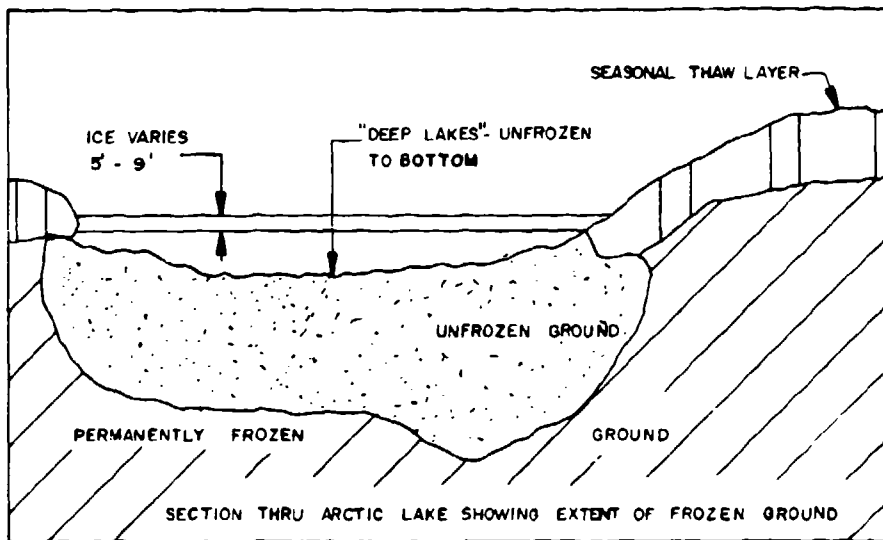
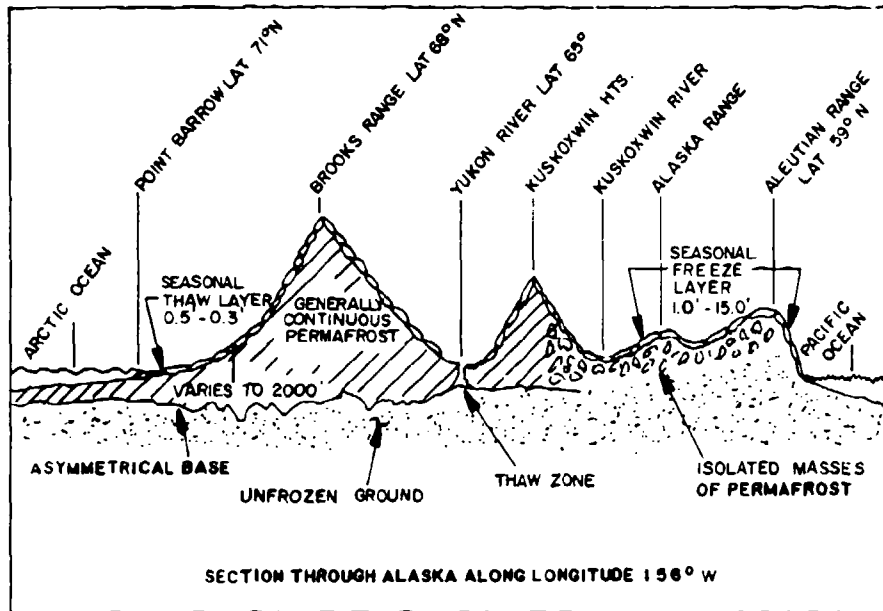


Figure 2-30. Illustration Showing Approximate Variations in Substructure

**2.11.2 Improving Electrical Grounding in Frozen Soils.** High electrical resistance of grounding sites is common in areas where the ground freezes. The performance of grounding installations can, however, often be increased through site selection and various electrode installation schemes. The degree of improvement will depend on the local existence and accessibility of conductive soils. The most common conductive sites are associated with thaw zones or clay-rich soils. The greatest grounding problems usually occur where bedrock, coarse-grained soil, or cold, ice-rich soil is found near the surface.

In temperate regions, small field installations can usually be adequately grounded by driving a simple vertical electrode into the soil. This technique has been unsuccessful in areas of frozen ground because: (1) driving electrodes is difficult, (2) frozen materials tend to be electrically resistive, and (3) high contact potentials can develop between a rod and the frozen soil because a thin ice layer can form around the cold rod.

Installation procedures can be modified in some frozen ground settings to eliminate some of these problems, permitting order-of-magnitude reductions in the resistance to ground. However, in many regions of the Arctic, electrical resistivity of the frozen ground is extremely high, and grounding may not be significantly improved by local modification or treatment of the soil surrounding the electrode. Achieving "low" resistance grounds of less than several ohms will often require that the site be selected in a zone of conductive material and is described in paragraph 2.11.1.

Other factors such as accessibility to water, power, roads, real estate, siting requirements, electromagnetic compatibility, etc, may however require that a site be located in an area of low soil conductivity. This establishes the rather high probability of not being able to attain a low resistance to ground without considerable cost and effort. Studies (2-17) conducted to determine methods to obtain low or acceptable resistances in areas of low soil conductivity in turn raised additional questions:

- a. What is the influence of ground temperature, material type and associated variations in unfrozen water content on the performance of an installation?
- b. What is the influence of material type and associated differences in permeability and saturation on salt solutions added to the soil surrounding an electrode?
- c. What is the effectiveness of using more than one electrode for lowering resistance to ground?
- d. What is the long-term influence of conductive backfills and what is the suitability of various materials for backfill around electrodes placed in holes of larger diameter than the electrodes?

The main procedure which can be used to reduce resistances to ground is to place the ground rod or electrode in open holes having diameters greater than the electrodes thereby making emplacement easier and permitting the use of conductive backfill. The holes can be made by drilling or blasting with shaped charges. Another procedure which may be used in limited situations is to lay or drive an array of horizontal rods into an active layer.

2.11.2.1 Electrode Resistance. The resistance of ground,  $R$ , of a single vertical electrode of length  $\ell$  in cm, and radius  $\alpha$  in cm, emplaced in homogeneous soil of resistivity  $\rho$  (ohm - cm) is found from:

$$R \text{ (ohms)} = \frac{\rho}{2\pi\ell} \left[ \ln \frac{4\ell}{\alpha} - 1 \right] \quad (2-77)$$

This equation may be used to estimate the penetration depth of conductive salt solutions in the soil adjacent to the treated backfill. Since the backfill is conductive, the electrode radius therefore is not just that of the metallic electrode, but initially the diameter of the hole filled with treated backfill. This large composite electrode is referred to as the effective electrode. For a constant ground temperature, any reduction in electrode resistance of a frozen saturated soil with time should be related to an increase in effective electrode diameter, presumably through salt movement. This increase can be determined by the soil resistivity  $\rho$  from equation 2-77 using the resistance to ground of the test electrode and the effective electrode radius measured at the time of installation. Periodically, after installation, the resistance to ground should be remeasured and the effective electrode radius can be calculated using the following form of equation 2-77 and using the soil resistivity calculated earlier:

$$\alpha \text{ (cm)} = \frac{4\ell}{\exp[1 + (2\pi\ell R/\rho)]} \quad (2-78)$$

#### 2.11.2.2 Installation and Measurement Methods.

2.11.2.2.1 Electrode Installation. Holes can be drilled with augers designed for use in frozen ground with hole diameters ranging from 3.8 cm (1-1/2 in.) to 91.4 cm (3 ft) and depths seldom greater than 2 m (6 ft). Hand-held equipment, consisting of an electric drive or a 5-hp gasoline-powered drill can also be used for most of the shallow, smaller-diameter holes. Both units could be used with a coring auger to drill holes up to 11 cm (4 in.) in diameter in fine-grained frozen soils. A truck-mounted auger can be used for the larger-diameter vertical holes drilled in coarse-grained materials. The horizontal electrodes can be hand-pushed and then driven into the thin seasonally thawed layer.

Military 6.8 kg (M2A3) shaped charges (used only by qualified personnel) can also be employed to produce vertical holes. Their similar performances in a range of frozen materials, with penetration approaching the length of standard electrodes, make this charge size ideal for electrode installation. The volume of several of the drilled holes can also be expanded by using C-4 block explosives.

2.11.2.2.2 Backfill. Reduction of contact potential is important in establishing a good electrical ground. In frozen soil, ice can form around the electrode, causing high contact resistance. Ice formation on the rod surface is likely since the rod is easily chilled by exposure of the upper end to low air temperatures. The beneficial effect of pouring untreated water around an electrode will only be short-term in cold environments. Therefore, the use of conductive backfill with a low freezing point becomes paramount to attain good ground or earth contact.

The backfill can be prepared by mixing salt and local soil or by saturating the soil backfill with a salt-water solution as shown in Figure 2-31. Backfill other than soil can also be used because soil is not always easily recovered from some drilled or blasted holes and because unfrozen material is difficult to find during the winter. Absorbent paper saturated with a salt solution and compacted in the hole around the electrode can also be used as a soil substitute.

The amount of salt added to the backfill is determined by preliminary laboratory conductivity measurements of several salt-soil mixtures. Salt may be added to silt and to a fine sand to obtain mixtures of from 0 to 20% salt based on the weight of the air-dried soil. Distilled water can be added to the salt-soil mixtures to obtain several soil moisture levels up to saturation for both materials. The soils should be compacted into a cylindrical plexiglass ring, which is clamped between electrodes for resistivity measurements at 1 kHz. Figure 2-32 shows the resistivity for two soils as a function of salt concentration at several volumetric moisture contents. A salt-soil mixture containing 1% salt results in a dramatic decrease in resistivity, with little effect after 5% salt for most moisture levels. Therefore, a 5% salt by weight is recommended for backfill as it produces a very conductive salt-soil mixture with the least amount of salt.

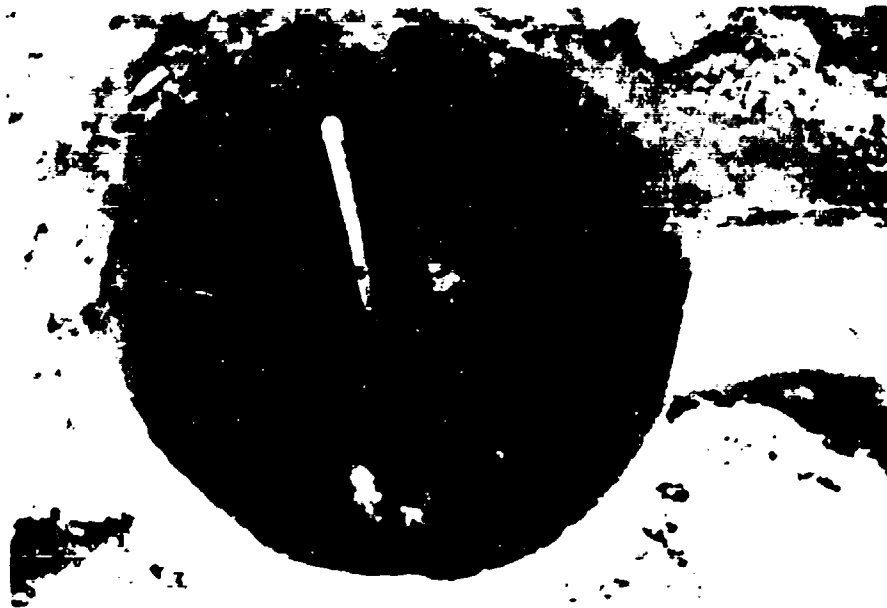


Figure 2-31. Installation of an Electrode During the Process of Backfilling with a Salt-Soil Mixture

Salt solution may also be poured around shallow-driven horizontal electrodes to minimize contact resistance during freezeback. These salt solutions in general may have concentrations on the order of 50-100%. Figure 2-33 shows a configuration of such horizontal electrodes placed in a thawed active layer.

Curves showing resistance-to-ground for metallic electrodes having various backfills are shown in Figures 2-34 through 2-38. Large seasonal variations are noted in electrode performance due to variations in unfrozen water content in both thawed and frozen materials. In some situations the improvement in grounding conditions during thaw periods can be extended by use of conductive backfill. The lower freezing point of the backfill will also reduce electrode contact resistance caused by freezing around the metallic electrodes.

Over a period of time, salt very likely will move into the soil adjacent to the electrode backfill and therefore will increase the effective area of the ground electrode and in turn reduce the resistance values. The level of the backfill should be checked annually to insure adequate levels are maintained to replenish this loss due to seepage.

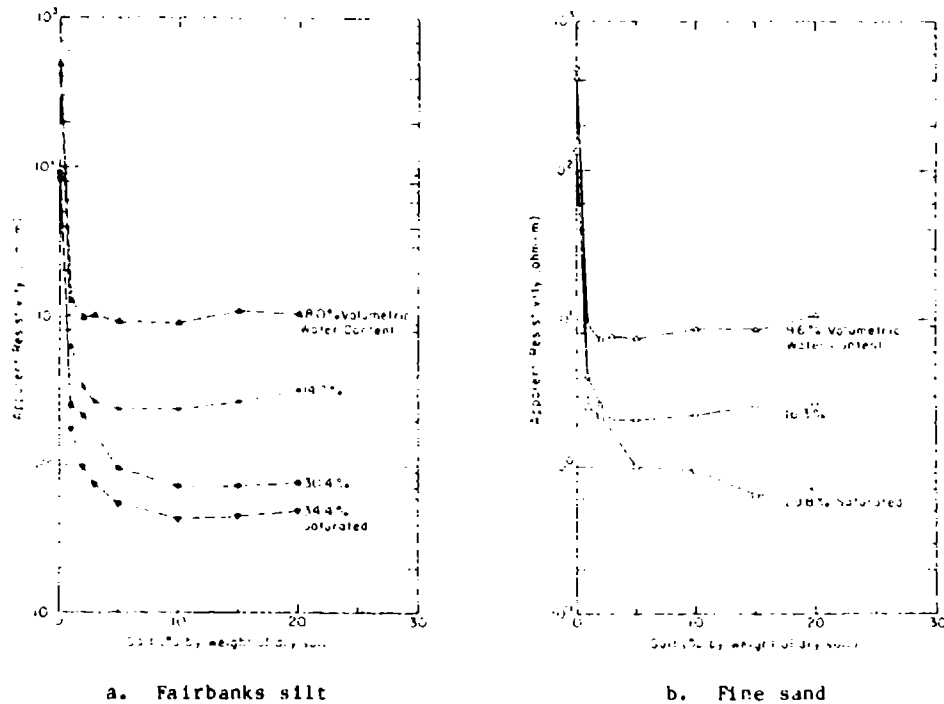


Figure 2-32. Apparent Resistivity for Two Soils at Various Moisture and Salt Contents

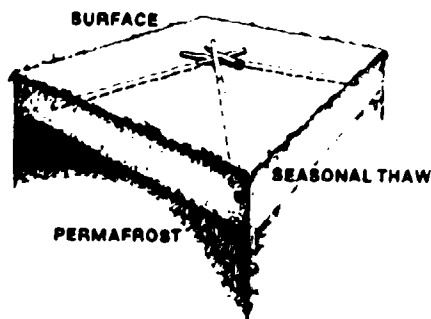


Figure 2-33. Configuration of Nearly Horizontal Electrodes Placed in the Thawed Active Layer

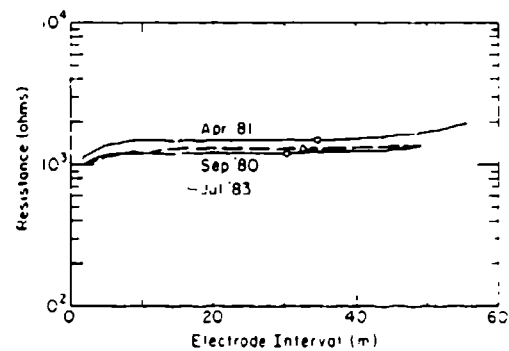


Figure 2-34. Resistance-to-Ground Curves for an Electrode Driven into Ice-Rich Silt

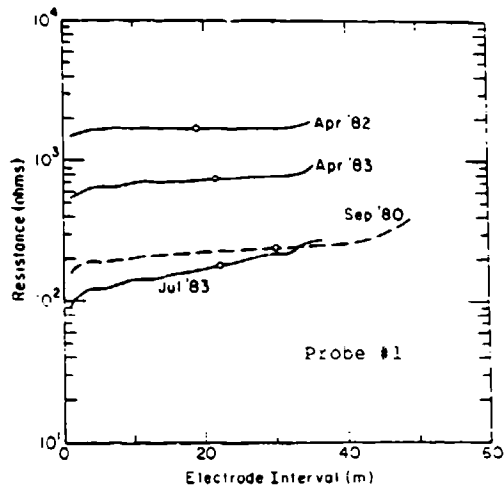


Figure 2-35. Resistance-to-Ground Curves for an Electrode Surrounded by a Backfill of Saturated Silt

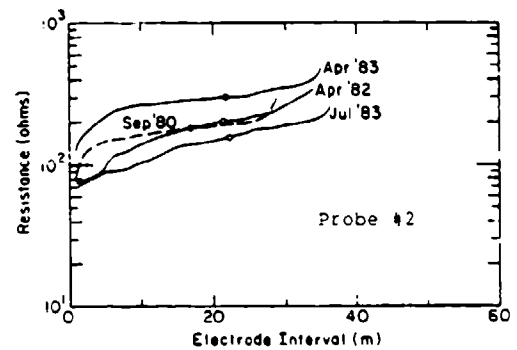


Figure 2-36. Resistance-to-Ground Curves for an Electrode Surrounded by a Water-Saturated Salt-Soil Backfill

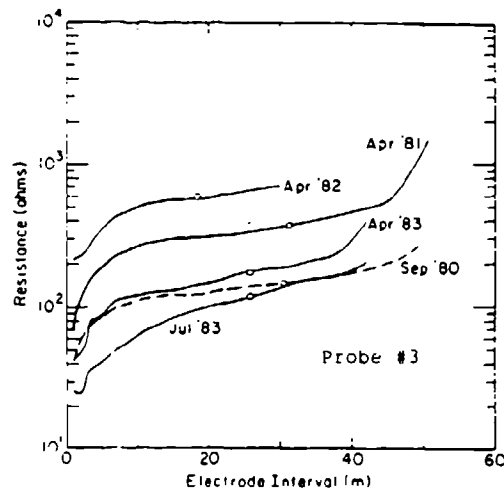


Figure 2-37. Resistance-to-Ground Curves for an Electrode Surrounded by a Water-Saturated Salt-Soil Backfill

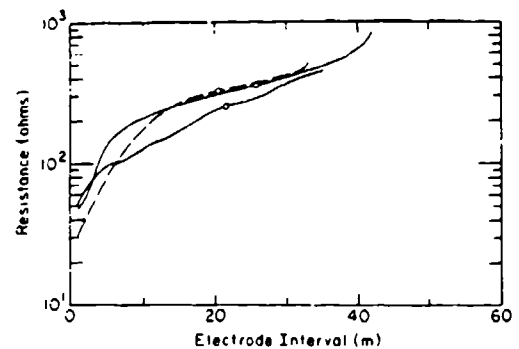


Figure 2-38. Resistance-to-Ground Curves for 3 Electrodes Placed in Holes Modified by Spring Charges and Filled with a Salt-Water Solution.



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## CHAPTER 3

## LIGHTNING PROTECTION SUBSYSTEM

3.1 THE PHENOMENON OF LIGHTNING.

Cumulonimbus clouds associated with thunderstorms are huge, turbulent air masses extending as high as 15 to 20 kilometers (9 to 12 miles) into the upper atmosphere. Through some means, not clearly understood, these air masses generate regions of intense static charge. These charged regions develop electric field gradients of hundreds, or perhaps thousands, of millions of volts between them. When the electric field strength exceeds the breakdown dielectric of air ( $\approx 3 \times 10^6$  volts/meter), a lightning flash occurs and the charged areas are neutralized.

Electric field measurements indicate that the typical thundercloud is charged in the manner illustrated by Figure 3-1 (3-1). A strong, negatively charged region exists in the lower part of the cloud with a counterbalancing positive charge region in the upper part of the cloud. In addition to these major charge centers, a smaller, positively charged region exists near the bottom of the cloud. Due to the strong negative charge concentration in the lower portion of the cloud, the cloud appears to be negatively charged with respect to earth -- except in the immediate vicinity underneath the smaller positive charge concentration.

Breakdown can occur between the charged regions within the cloud to produce intracloud lightning. It can also occur between the charged regions of separate clouds to produce cloud-to-cloud lightning. Intracloud and cloud-to-cloud discharges do not present a direct threat to personnel or structures on the ground and thus tend to be ignored in the design and implementation of lightning protection systems. However, calculations of the voltages which could be induced in cross-country cables by such discharges (3-2) indicate that they present a definite threat to signal and control equipments, particularly those employing solid state devices.

The cloud-to-ground flash is the one of primary interest to ground-based installations. By definition, such flashes take place between a charge center in the cloud and a point on the earth. This point on earth can be a flat plain, body of water, mountain peak, tree, flag pole, power line, residential dwelling, radar or communications tower, air traffic control tower, or multi-story skyscraper. In a given area, certain structures or objects are more likely to be struck by lightning than others; however, no object whether man-made or natural feature, should be assumed to be immune from lightning.

The high currents which flow during the charge equalization process of a lightning flash can melt conductors, ignite fires through the generation of sparks or the heating of metals, damage or destroy components or equipments through burning or voltage stressing, and produce voltages well in excess of the lethal limit for people and animals. The objective of all lightning protection subsystems is to direct these high currents away from susceptible elements or limit the voltage gradients developed by the high currents to safe levels.

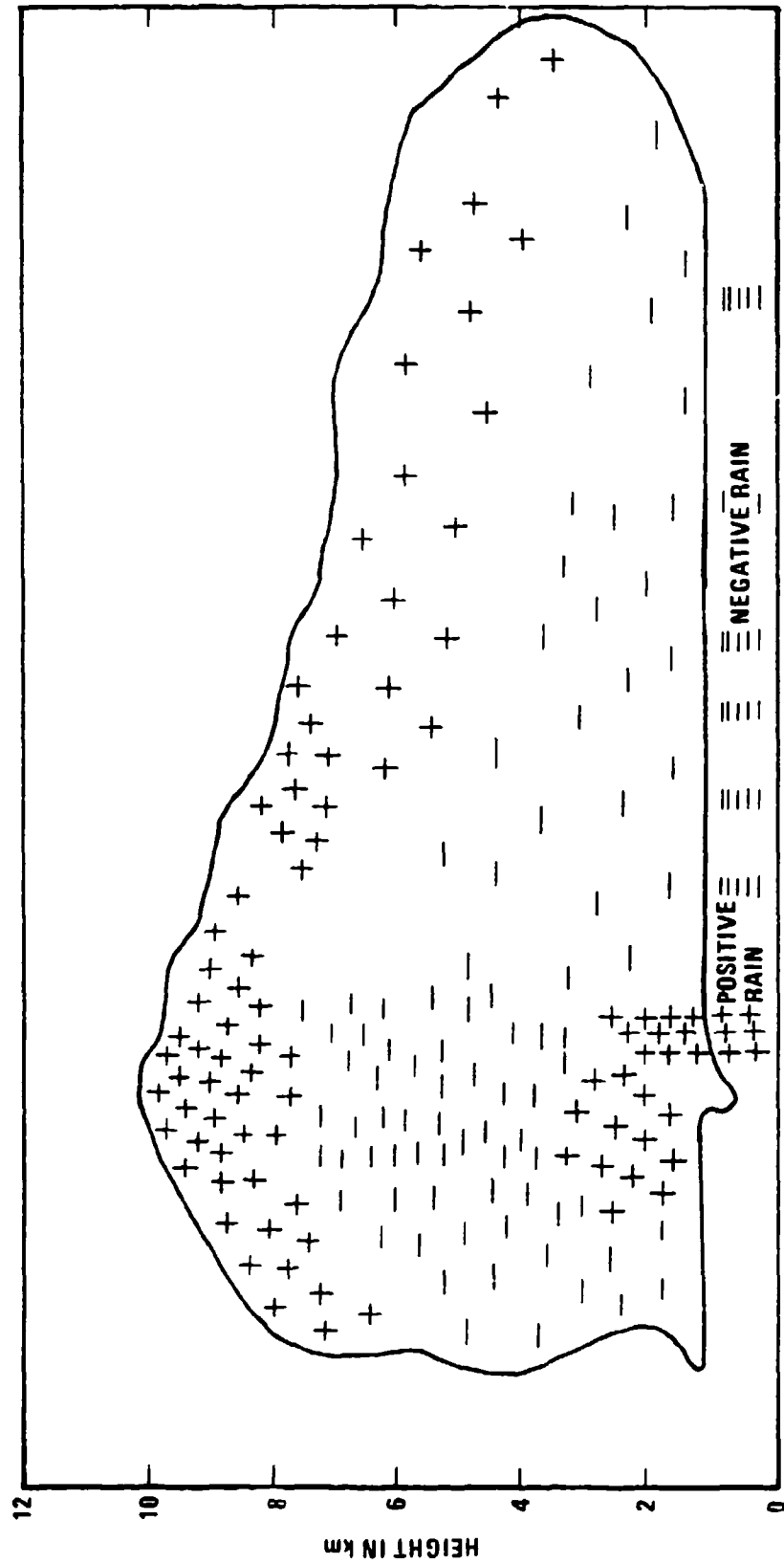


Figure 3-1. Charge Distribution in a Thundercloud

### 3.2 DEVELOPMENT OF A LIGHTNING FLASH.

As the charge builds up in a cloud, the electric field in the vicinity of the charge center builds up to the point where the air starts to ionize. A column of ionized air, called a pilot streamer, begins to extend toward earth at a velocity of about 160 kilometers per hour (100 miles per hour) (3-3). After the pilot streamer has moved perhaps about 30 to 45 meters (100 feet to 150 feet), a more intense discharge called a stepped leader takes place. This discharge lowers additional negative charge into the region around the pilot streamer and allows the pilot streamer to advance for another 30 to 45 meters (100 to 150 feet) after which the cycle repeats. The stepped leader progresses towards the earth in a series of steps with a time interval between steps on the order of 50 microseconds (3-4).

In a cloud-to-ground flash, the pilot streamer does not move in a direct line towards the earth but instead follows the path through the air that ionizes most readily. Although the general direction is toward the earth, the specific angle of departure from the tip of the previous streamer that the succeeding pilot streamer takes is rather unpredictable. Therefore, each 30 to 45 meter (100 to 150 foot) segment of the discharge will likely approach the earth at a different angle. This changing angle of approach gives the overall flash its characteristic zig-zag appearance.

Being a highly ionized column, the stepped leader is at essentially the same potential as the charged area from which it originates. Thus, as the stepped leader approaches the earth, the voltage gradient between the earth and the tip of the leader increases. The increasing voltage further encourages the air between the two to break down.

The final stepped leader bridges the gap between the downward progressing column and the earth or an extension of the earth such as a tree, building, or metal structure that is equipotential with the earth. While the stepped leader is approaching the earth, a positive charge equivalent to the negative charge in the cloud is accumulating in the general region underneath the approaching leader. Once the stepped leader contacts earth (or one of its extensions), the built-up positive charge in the earth flows rapidly upward through the ionized column established by the stepped leader to neutralize the strong negative charge of the cloud. This return current constitutes what is generally referred to as the lightning stroke. If additional pockets of charge exist in the cloud, these pockets may discharge through the ionized path established by the initial stroke. Continuous dart leaders proceed from a remaining charge pocket toward the earth down this path. Once the dart leader reaches the earth, another return stroke of positive charge propagates up the channel to neutralize the secondary charge in the cloud. This cycle may be repeated several times as succeeding charge centers in the cloud are neutralized.

### 3.3 INFLUENCE OF STRUCTURE HEIGHT.

Flashes to earth are normally initiated by a pilot streamer from the cloud. As the charged leader approaches the ground, the voltage gradient at the surface increases. Ultimately the voltage becomes high enough for an upward-moving leader to be induced. Over flat, open terrain, the length of the upward leader does not exceed a few meters before it unites with the downward leader to start the return stroke. However, structures or other extensions from the earth's surface experience intensified electric field concentrations at their tips. Consequently the upward leaders are generated while the downward leader is some distance away; the upward

leader can be several hundred meters long before the two meet. For very tall buildings, the upward leaders begin to form even before the downward leaders have begun to form within the cloud; such incidents are generally described as triggered lightning. Triggered lightning is not very common for structures less than 150 meters (500 feet) in height; as the height increases above this threshold, the proportion of triggered strikes increases rapidly (3-5).

#### 3.4 STRIKE LIKELIHOOD.

The number of total flashes to which the structure is exposed is related principally to local thunderstorm activity. Local thunderstorm activity can be projected from isokeraunic maps similar to those shown in Figures 3-2 and 3-3. These maps show the number of thunderstorm days per year for various regions of the United States and the world. Additional maps of worldwide keraunic levels can be obtained from the World Meteorological Association (3-6).

A thunderstorm day is defined as a local calendar day on which thunder is heard irrespective of whether the lightning flashes are nearby or at some distance away. To an observer at a specific location, the average distance at which lightning may occur and thunder will be heard is about 10 km (6 miles) (3-5). Therefore, a thunderstorm day means that at least one lightning discharge has occurred within an area of about 300 square km (120 square miles) surrounding the position of the observer. The actual number of strikes in the immediate vicinity of the observer may be considerably higher or lower than the number of thunderstorm days might indicate, depending upon the duration and intensity of a specific storm or series of storms.

In spite of the relative inexactness of a prediction of a lightning strike to a specific object that is based on the keraunic level, the thunderstorm day is the only parameter related to lightning incidence that has been documented extensively over many years. Its primary value lies in the qualitative information which it provides. This information can be used to assist in the determination of whether lightning protection should be provided in those situations where there is serious doubt as to the relative need for such protection. For example, a particular facility may not be essential to the safety of aircraft, but the loss of the facility may cause traffic delay. In an area of frequent thunderstorms such as the west coast of Florida, for example, the number of outages in areas where there was no protection could be so high as to be unacceptable; in an area of few thunderstorms; e.g., Southern California or Alaska, the expected outage from lightning might be once every few years (which could be significantly less than outages for routine maintenance).

The number of lightning flashes per unit earth surface area increases with the number of thunderstorm days per year, though not linearly. Empirical evidence indicates that the number of flashes per square kilometer,  $\sigma_y$ , can be reasonably predicted from (3-5):

$$\sigma_y = 0.007 T_y^2 \quad (3-1)$$

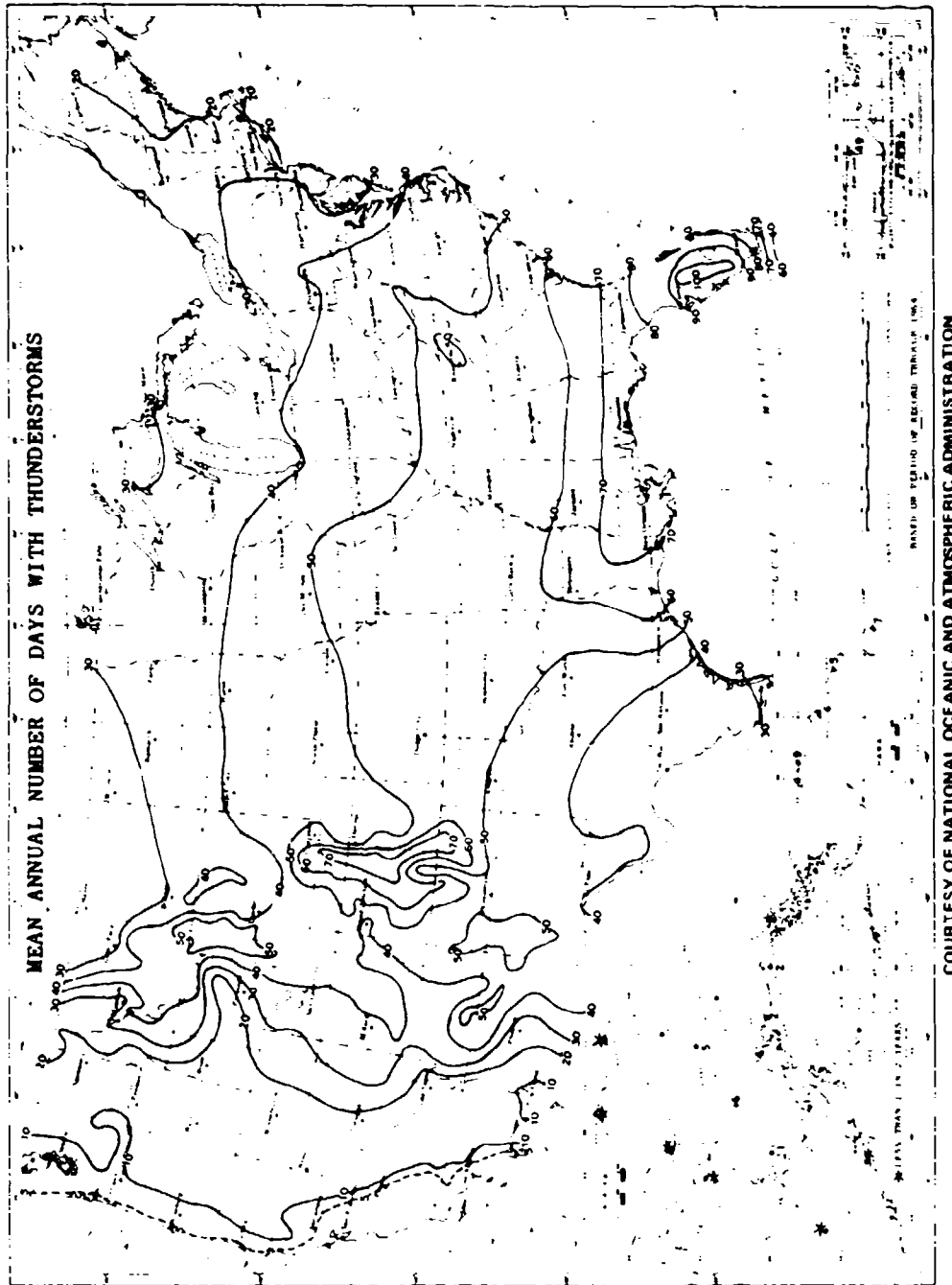
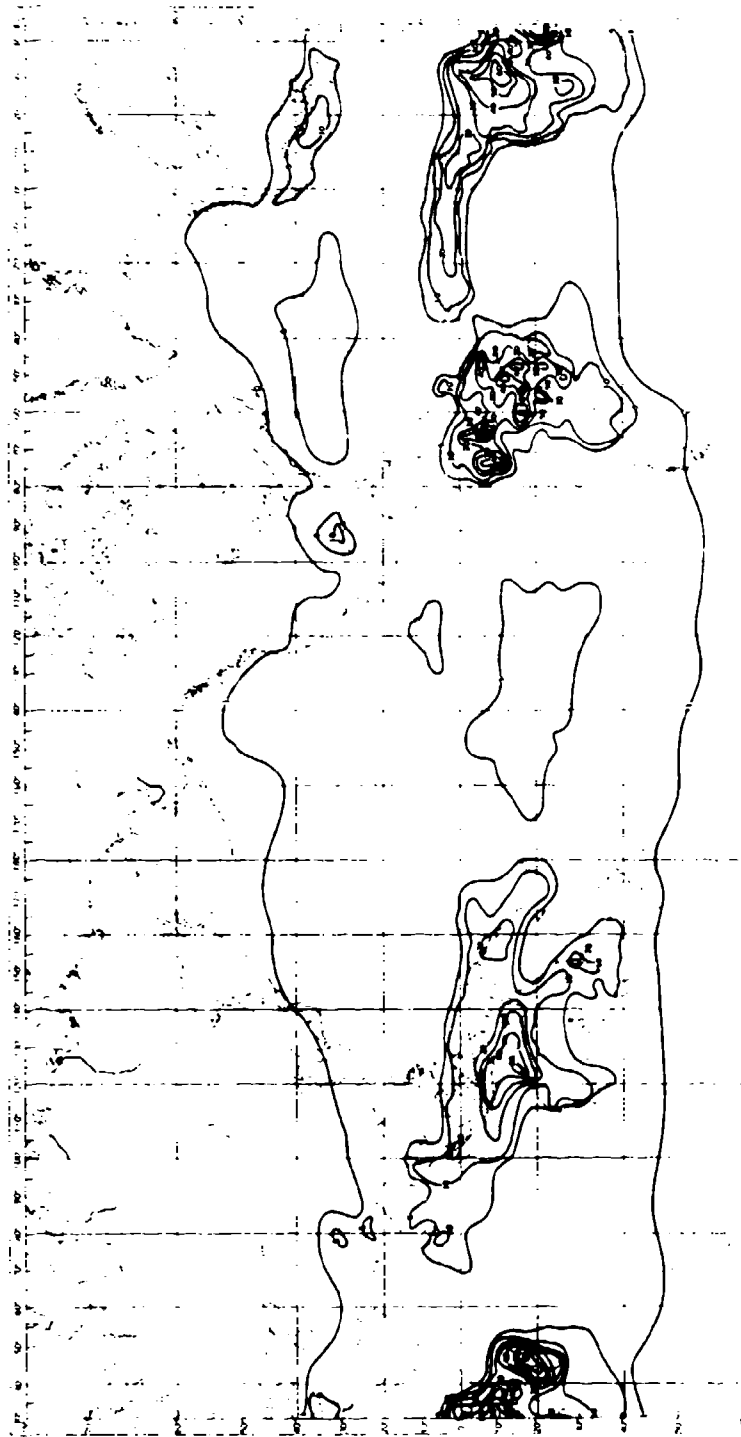


Figure 3-2. Mean Number of Thunderstorm Days Per Year for the United States



COURTESY OF  
NATIONAL OCEANIC AND  
ATMOSPHERIC ADMINISTRATION

DEC - JAN - FEB Average number of days with THUNDERSTORMS

Figure 3-3. Worldwide Isokeraunic Map (Sheet 1 of 4)

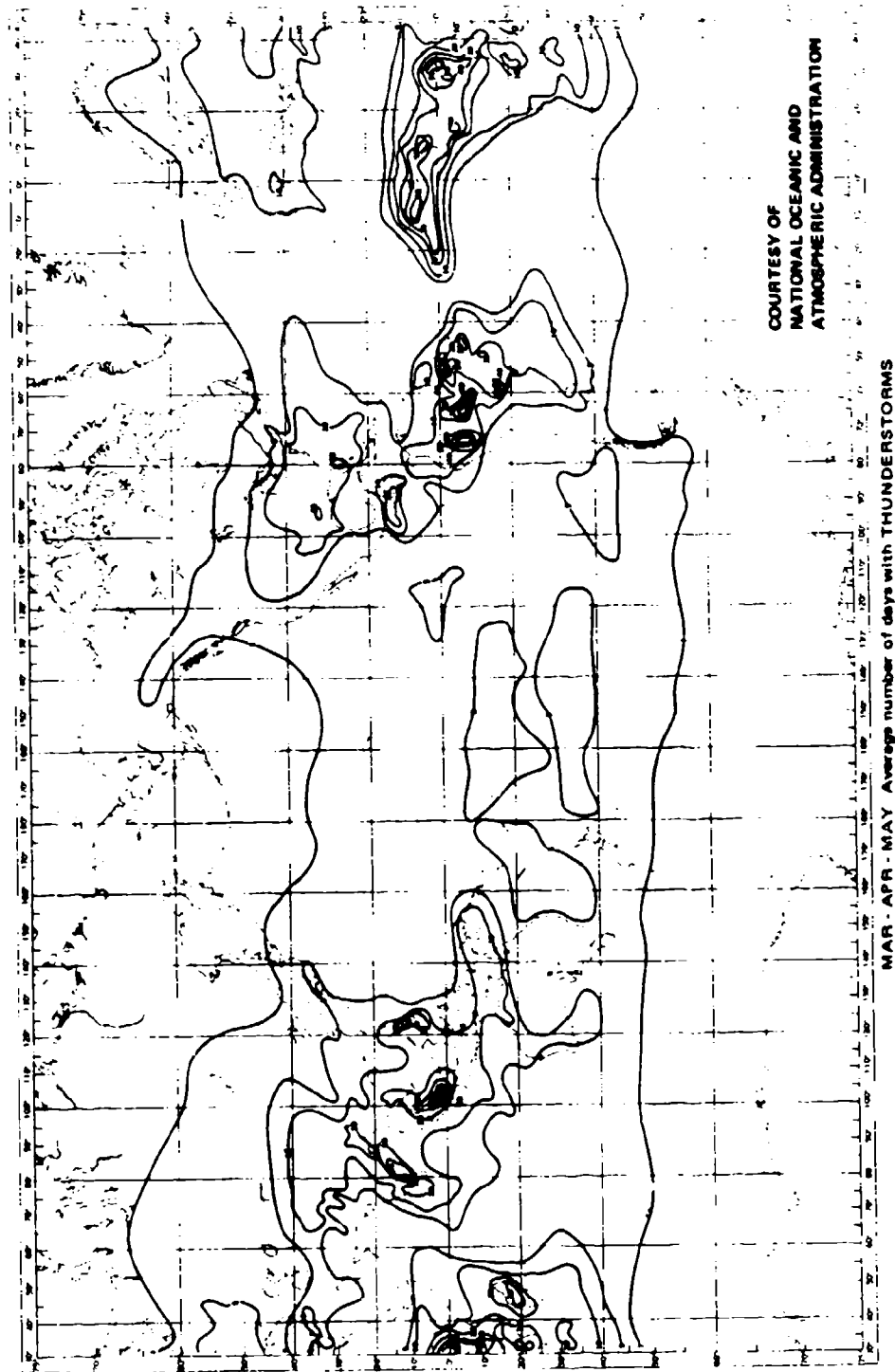


Figure 3-3. Worldwide Isothermic Map (Sheet 2 of 4)



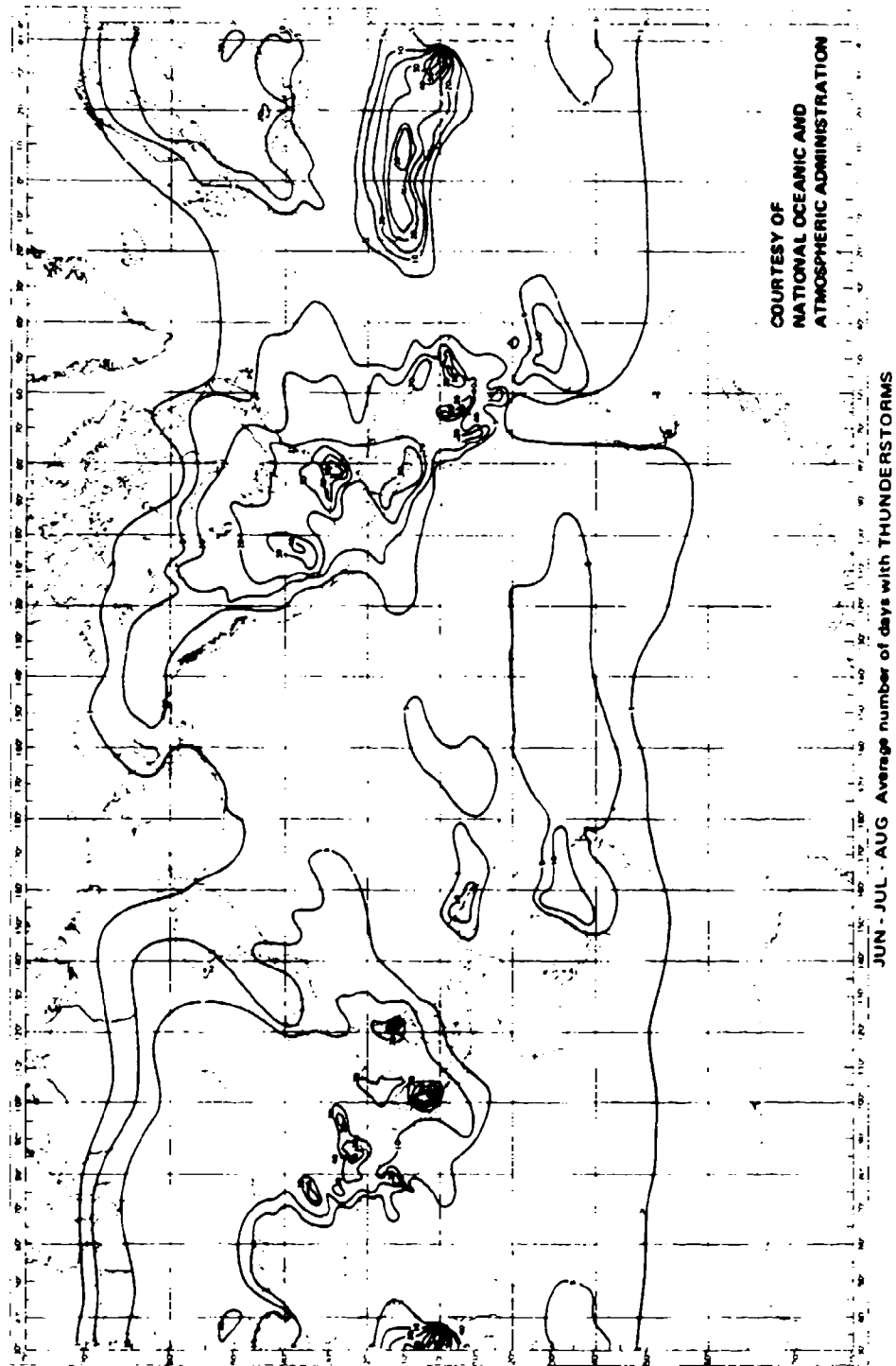


Figure 3-3. Worldwide Isokeraunic Map (Sheet 3 of 4)

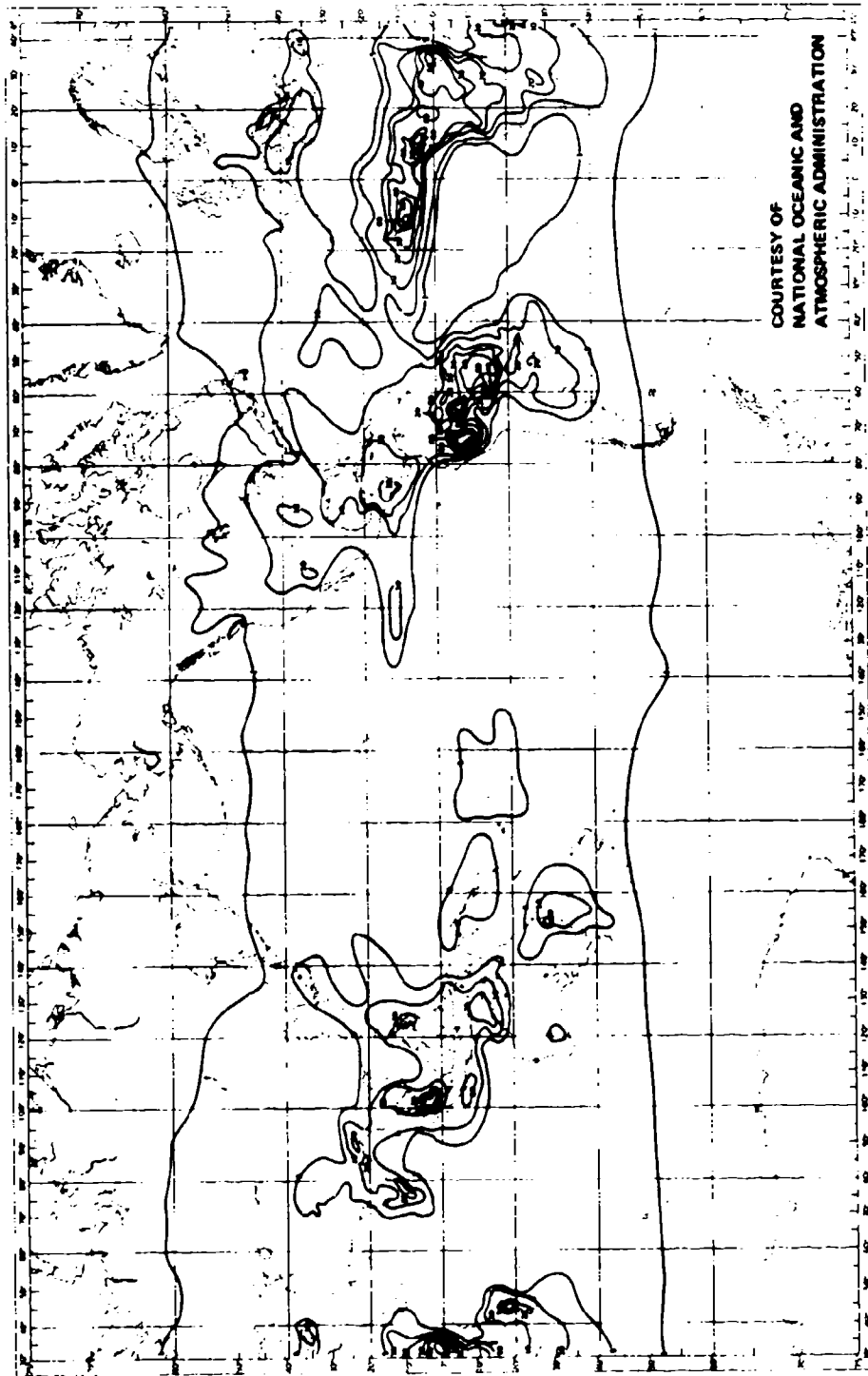


Figure 3-3. Worldwide Isokeraunic Map (Sheet 4 of 4)

where  $T_y$  is the number of thunderstorm days per year. Out of the total number of flashes per unit area, the number of discharges increases with increasing geographical latitude (3-7). The proportion,  $p$ , of discharges that go to ground in relation to the geographical latitude,  $\Lambda$ , can be represented (3-8) as:

$$p = 0.1 (1 + (\Lambda/30)^2) \quad (3-2)$$

Thus in a given location the flash density,  $\sigma_{yg}$ , i.e., the number of discharges to earth per square kilometer per year, is:

$$\sigma_{yg} = p\sigma_y \quad (3-3)$$

To calculate  $\sigma_{yg}$ , for a specific location, first determine  $T_y$  from the isokeraunic map of Figure 3-2 to Figure 3-3. For estimation purposes, the number of thunderstorm days at points between lines may be determined by interpolation. Using this value of  $T_y$ , calculate the total flash density with Equation 3-1. Next obtain the geographical latitude of the site from a map of the area and calculate  $p$  from Equation 3-2. Then determine the number of strikes to earth per year per square kilometer with Equation 3-3.

**3.5 ATTRACTIVE AREA.** The concept of attractive area reflects the principle that an object extending above its surroundings is more likely to be struck by lightning than its actual cross-sectional area might otherwise indicate. For example, thin metallic structures such as flag poles, lighting towers, antennas, and overhead wires offer a very small cross-sectional area relative to the surrounding terrain but ample evidence exists to show that such objects apparently attract lightning.

#### 3.5.1 Structures Less Than 100 Meters High.

For structures less than 100 meters (330 feet) in height, and which therefore do not normally trigger lightning, the number of strikes increases according to a power of  $h$ , the structure height. An expression that represents the attractive radius,  $r_a$ , in meters of a structure is (3-5);

$$r_a = 80 \sqrt{h} (e^{-0.02h} - e^{-0.05h}) + 400 (1 - e^{-0.0001h^2}) \quad (3-4)$$

where  $h$  is in meters. For a structure 10 meters high, Equation 3-4 gives an attractive radius of 57.7 meters; similarly, the attractive radius for a 100-meter high structure is 356 meters. The attractive area,  $A_a$ , is  $\pi r_a^2$ . Thus  $A_a$  for a 10-meter structure is approximately 0.01 square kilometer, while the attractive area of a 100-meter structure is 0.4 square kilometer.

Equation 3-4 has been found to adequately describe the number of strikes to objects which are not tall enough to trigger lightning. For taller structures, a multiplication factor (3-5)

$$F_T = 1 + 2 (9-1500/h) \quad (h \text{ in meters}) \quad (3-5)$$

should be applied to Equation 3-4. The experimental data to justify the use of Equation 3-5 for structures greater than 400 meters (1300 feet) is sketchy. However, since structures even approaching this height are not expected to be of primary concern, Equations 3-4 and 3-5 are expected to be adequate for most design purposes.

Large flat buildings that do not extend above the median treetop level in the general area will have an attractive area that is essentially the area of the roof (assuming the roof covers the entire structure). If the building is several stories high such that it appreciably extends above the prevailing terrain, then its attractive area is its roof area plus that portion of the attractive area not already encompassed by the roof. Figure 3-4 illustrates the method for calculating the attractive area of a rectangular structure of length,  $l$ , and width,  $w$ . The roof area is given by  $l \times w$ . The additional attractive area resulting from the height of the building is readily determined by recognizing that the areas contributed by the four corners of the building equal a circle of radius,  $r_a$ . Both ends of the structure (dimension  $w$ ) contribute the area of  $2 \pi r_a^2$ ; the sides contribute  $2l \pi r_a$ . The total attractive area is the sum of the roof area ( $lw$ ), the corners ( $\pi r_a^2$ ), the ends ( $2 \pi r_a^2$ ), and the sides ( $2l \pi r_a$ ) to produce a total of

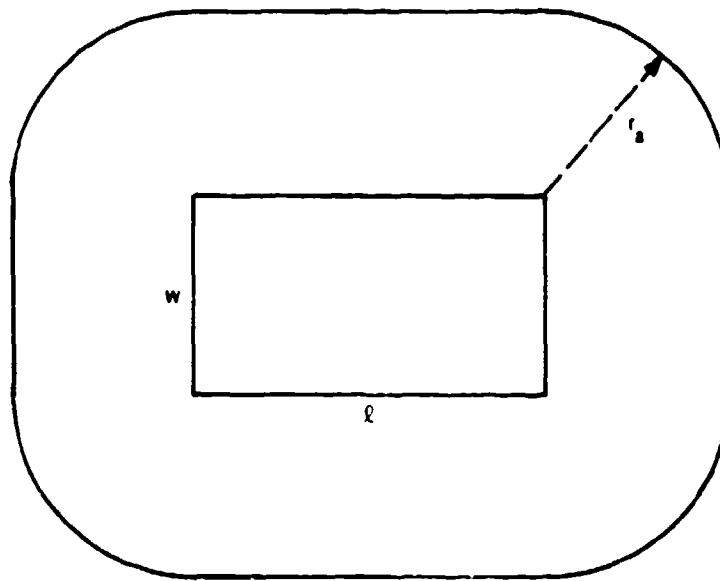
$$A_a = lw + \pi r_a^2 + 2 \pi r_a (w + l). \quad (3-6)$$

Figure 3-5 indicates that the height to be used in calculating the attractive area of a tall structure should be the height that the structure extends above the effective (i.e., the level that earth charges would rise to if the building were not there) levels of the earth. On open, level terrain the height,  $h$ , would be the full height of the roof from grade level.

The number of flashes which can be expected to strike a given structure is equal to the product of the flash density,  $\sigma_{yg}$ , times the attractive area,  $A_a$ , of the structure. For example, suppose the relative likelihood of a lightning strike to a low, flat structure 100 meters on a side, located in Nashville, TN, is desired. From Figure 3-2,  $T_y$  is determined to be approximately 54 thunderstorm days per year. The flash density as given by Equation 3-1 is 20.4 flashes/km<sup>2</sup>/year. The proportion of those flashes that are discharges to earth is 24.4 percent (from Equation 3-2) since the latitude is 36 degrees. Thus approximately 5 flashes/km<sup>2</sup>/year to earth can be expected. Within the area of the structure (0.01 km<sup>2</sup>) there will be only 0.05 strikes per year on the average, or there is a 1 in 20 chance of being struck by lightning in a given year. For the same structure in Southern California, only a 1 in 330 likelihood of a strike would be expected in a given year.

### 3.5.2 Cone of Protection.

This ability of tall structures or objects to attract lightning to themselves serves to protect shorter objects and structures. In effect, a taller object establishes a protected zone around it. With this protected zone, other shorter structures and objects are protected against direct lightning strikes. As the heights of these shorter objects increase, the degree of protection decreases. Likewise, as the separation between tall and short structures increases, the protection afforded by the tall structure decreases. The protected space surrounding a lightning conductor is called the zone (or cone) of protection.



TOTAL ATTRACTIVE AREA:  $A_a = wl + \pi r_a^2 + 2r_a(w + l)$

$r_a$  IS DETERMINED BY EFFECTIVE HEIGHT,  $h$ . SEE EQUATIONS 3-4 AND 3-5.

Figure 3-4. Attractive Area of a Rectangular Structure

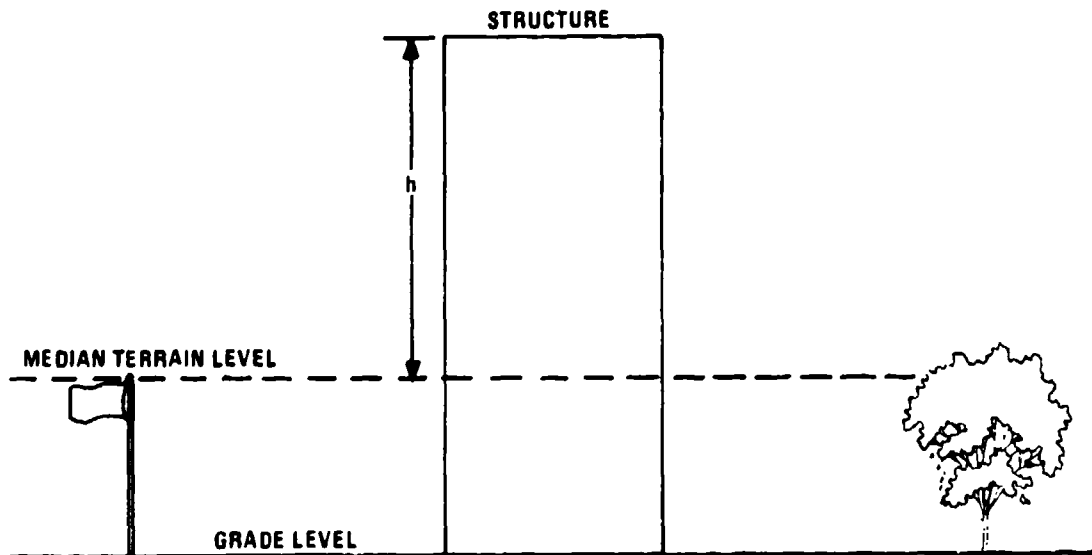


Figure 3-5. Effective Height of a Structure

The zone of protection provided by a grounded vertical rod or mast is conventionally defined as the space enclosed by a right circular cone with its axis coincident with the mast and its apex at the top of the mast as illustrated by Figure 3-6(a). Similarly, the zone protected by a grounded horizontal overhead wire is defined as a triangular prism with its upper edge along the wire as illustrated in Figure 3-6(b). In either case, the zone (or cone) of protection is expressed as the ratio of the horizontal protected distance,  $D$ , to height,  $H$ , of the mast or wire. This ratio is the tangent of the shielding angle,  $\alpha^*$ . Some commonly recommended zones of protection and the associated shielding angles are illustrated in Figure 3-7.

The NFPA Lightning Protection Code (3-9) recommends that a 1:1 zone of protection ( $\alpha = 45^\circ$ ) be provided in important areas while a 2:1 zone ( $\alpha = 63^\circ$ ) is acceptable for less important areas. The British Standard Code of Practice (3-10) states that a shielding angle of 45 degrees provides an acceptable degree of protection for ordinary structures, but that for structures with explosive or high flammable contents the shielding angle should not exceed 30 degrees.

Although the existence of a 1:1 zone of protection does not absolutely guarantee immunity to lightning, documented cases of the 1:1 zone being violated are very few. Thus for all facilities except those associated with the storage of explosives or fuels, a 1:1 zone of protection can safely be used as a basis of design of lightning protection systems. As such, C-E facilities or equipments (antennas, etc.) located entirely within the 1:1 zone of protection generally are not required to have separate air terminals. This does not eliminate the need to ground metal shelters or to meet the grounding requirements of the subsystems which comprise the facility ground system. If more than one rod or wire is used, the protected zone is somewhat greater than the total of all of the 1:1 zones of the rods or wires considered individually. For adjacent structures, the Codes specify that a 2:1 zone of protection may be assumed for the region between the structures.

Large structures with flat or gently sloping roofs do not lend themselves to the straightforward application of the 1:1 or 2:1 zone of protection principles. To establish even 2:1 type coverage on large buildings, exceptionally tall air terminals would be required. Experience, however, shows that extremely tall terminals are not needed for effective protection. Both the NFPA Lightning Protection Code and UL Master Labeled Protection System (3-11) specify air terminals that extend from 10 to 36 inches above the object to be protected. (The British Standard Code of Practice does not require the use of air terminals at all.)

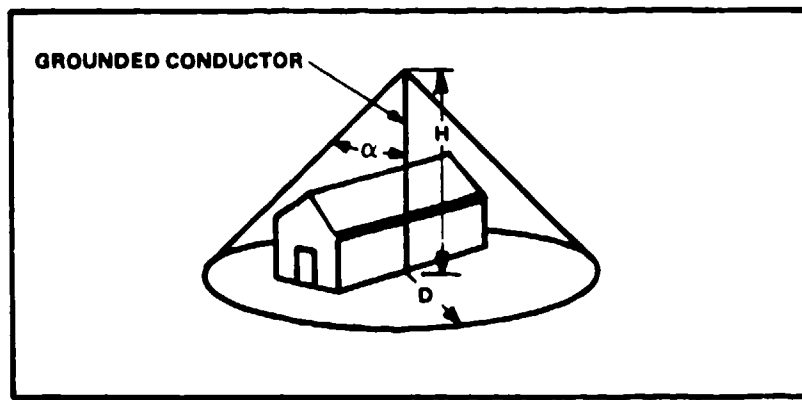
### 3.6 LIGHTNING EFFECTS.

#### 3.6.1 Flash Parameters.

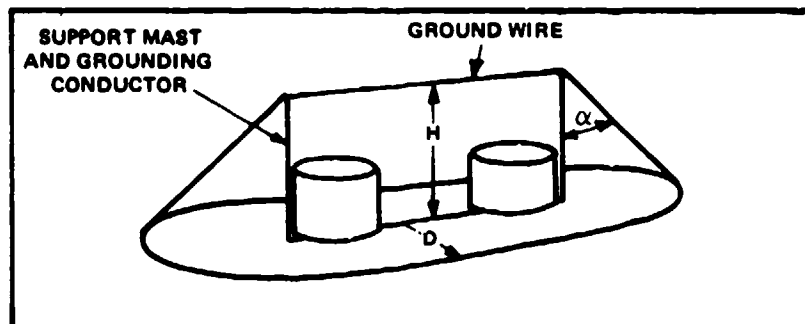
During the short interval of a lightning flash, several discharges occur. The sequence of events in a multiple-stroke flash is illustrated in Figure 3-8. The initial path for the discharge is established in 50 microseconds. Intermediate return stroke currents of about 1 kA follow the initial return stroke and last for a few milliseconds. Subsequent strokes occur at intervals of 50 to 60 milliseconds. The return stroke interval may include a continuing current of 100 A or so which flows for several milliseconds or until the start of the next return stroke.

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\*The shielding angle is defined as the angle between the surface of the cone and a vertical line through the apex of the cone, or between the side of the prism and the vertical plane containing the horizontal wire.



(a) CONE OF PROTECTION PROVIDED BY A VERTICAL GROUND CONDUCTOR OF HEIGHT H.



(b) ZONE OF PROTECTION PROVIDED BY A HORIZONTAL AERIAL GROUND WIRE AT HEIGHT H.

Figure 3-6. Zones of Protection Established by a Vertical Mast and a Horizontal Wire

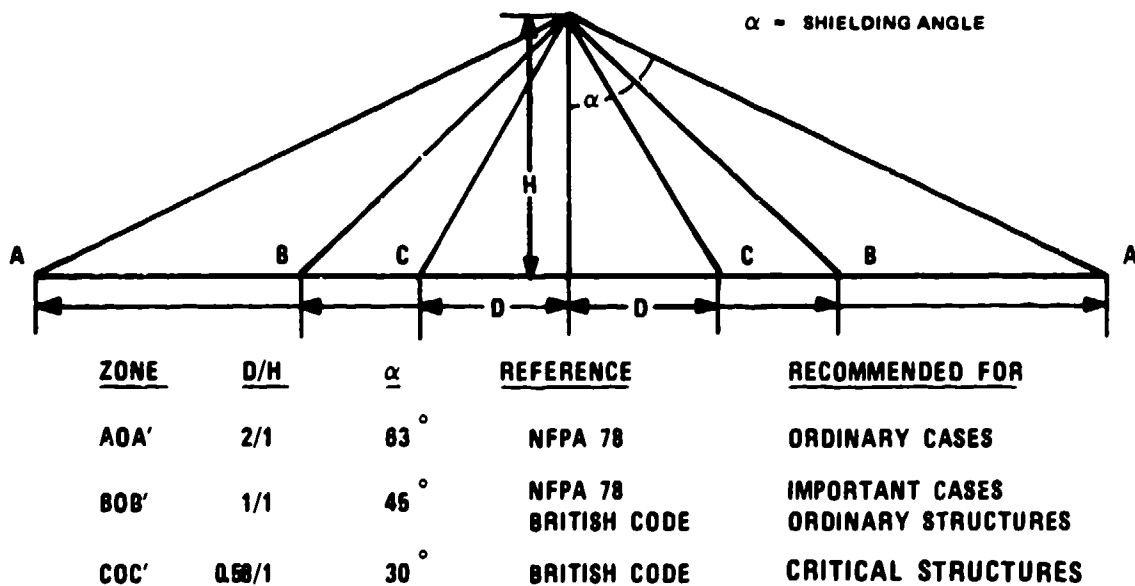


Figure 3-7. Some Commonly Used Lightning Shielding Angles

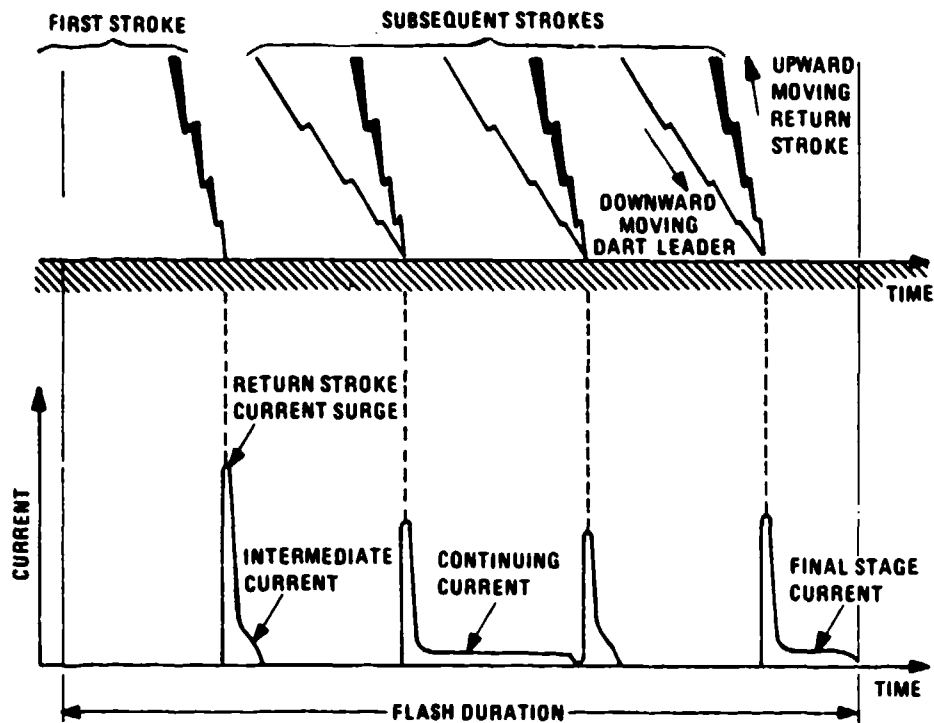


Figure 3.8. Illustration of Process and Currents Which Occur During a Lightning Flash to Ground

The lightning discharge involves the transfer of large amounts of electric charge between the cloud and the earth. The typical flash transfers 15 to 20 coulombs (C) (1 coulomb equals  $6.2 \times 10^{18}$  electrons) with some flashes involving as much as 400 coulombs of charge. The energy per flash of lightning has been estimated to be as high as  $10^8$  watt-seconds. Table 3-1 summarizes the range of values for selected lightning parameters.

### 3.6.2 Mechanical and Thermal Effects.

The fast rise time, high peak amplitude current of the stroke can produce severe mechanical, thermal, and electrical effects. The damage caused by these currents to objects in the discharge path is closely related to the relative conducting power of the object. For example, metals generally receive a discharge with little damage. In most cases, even slender conductors such as telephone and electric power cables handle the current without fusing (melting) except at the point where the current enters or leaves the metal (where severe damage may occur). Very strong discharges of high peak current ( $> 40$  kA) and high coulomb values ( $> 200$  C), however, can melt or burn holes in solid metal plates. This burning effect is not usually of primary concern for a typical building or structure because, if an adequate protection system is installed, the principle effect will be a small deformation at the tip of a lightning rod or a small melted area on the intercepting cable. Such effects are of more concern where flashes to airplanes occur because such burning can perforate the fuselage to cause loss of pressurization or penetrate the skin of fuel tanks and possibly ignite fuels. The burning or melting also presents a threat to exposed tanks of volatile gases or fuels on the ground.



Table 3-1

## Range of Values for Lightning Parameters (3-5)

<u>Parameter</u>	<u>Minimum</u>	<u>Typical</u>	<u>Maximum</u>
Number of return strokes per flash	1	2 to 4	26
Duration of flash(s)	0.03	0.2	2
Time between strokes (ms)	3	40 to 60	100
Peak current per return stroke (kA)	1	10 to 20	250
Charge per flash (C)	1	15 to 20	400
Time to peak current ( $\mu$ s)	<0.5	1.5 to 2	30
Rate of rise (kA/ $\mu$ s)	<1	20	210
Time to half-value ( $\mu$ s)	10	40 to 50	250
Duration of continuing current (ms)	50	150	500
Peak continuing current (A)	30	150	1600
Charge in continuing current (C)	3	25	330

Because of the duration of the currents that flow for the extended intervals between return strokes, they are most likely to cause damage by melting or igniting solid materials. In contrast, the short-duration high-current peaks tend to tear or bend metal parts by the electromagnetic forces that develop in proportion to the square of the instantaneous current. Though potentially hazardous, the damage caused by mechanical forces in metallic conductors is generally of secondary importance in most situations. However, because of the presence of these mechanical forces, it is necessary that lightning rods, down conductors, and other elements of the protection system be securely anchored.

On the other hand, when insulating or semi-insulating material receives a discharge, an explosive reaction may occur with severe damage. Trees, for instance, whether dry or green, are in many cases split or stripped of their bark, and the damage can extend underground to their roots. Related damage may occur to other unprotected wooden structures or objects such as flag poles, masts, or light supports, and electric and telephone poles. When lightning strikes a wooden building, the stroke seeks out the lowest impedance path to earth which is probably through the electric wiring or water pipes. Often in order to reach these metallic paths, the discharge must pass through some type of wooden barrier. In penetrating such barriers, extensive explosive damage usually results.

Brick, concrete, marble, and other masonry materials are also frequently shattered or broken loose at the point where the discharge passes through them. Such damage will occur where structural steel support members or steel reinforcing rods are encased in concrete or sheathed in brick or marble and the structure has an inadequate protective system. The explosive effect can dislodge materials with considerable force -- force sufficient to hurl relatively large pieces several meters. One explanation of the explosive force is that it is the result of the virtually instantaneous vaporization of the water present in the wood or entrapped in the masonry materials.

### 3.6.3 Electrical Effects.

Lightning discharges to or near the buildings and structures frequently cause damage to electrical and electronic equipment. Melting or burning of conductors occurs at the point of interception of the stroke. The voltages developed by the fast risetime, high amplitude current pulse are frequently high enough to break down insulation, pose personnel hazards, and cause component and device failure. These voltages are produced by:

- a.  $IZ$  (current  $\times$  impedance) drop resulting from the lightning pulse traveling down power lines or signal lines, through structural members, along down conductors or overhead ground wires or through the resistance of the earth connection;
- b. Magnetic induction; and
- c. Capacitive coupling.

Lightning surges in power, signal, and control circuits are generally the result of some combination of these three components.

#### 3.6.3.1 Conductor Impedance Effects.

Because of the fast risetime (1 to 2  $\mu\text{sec}$ ) and high amplitude (10 to 20 kA) characteristics of the current pulse produced by the lightning discharge, the inductance and resistance of even relatively short conductors causes extremely high voltages to be developed on the conductor. The voltages frequently are high enough to exceed the breakdown potential of air or other insulation materials and cause flashover to other conductors or breakdown of insulation. The resistive  $IR$  drop generated by 20 kA in a 30 meter (100 feet) run of down conductor conforming to NFPA-78 ( $2.88 \times 10^{-4} \Omega/\text{m}$ ) will be

$$V = 2 \times 10^4 \times 2.88 \times 10^{-4} \times 30 = 173 \text{ volts} \quad (3-7)$$

which is not sufficient to cause flashover or to pose a serious threat to personnel.

For a down conductor length of 30 meters (100 feet), the smallest copper conductor meeting the minimum requirements of the Lightning Protection Code or the UL Master Labeled Lightning Protection System has a diameter of 0.894 cm (0.352 inches). Assuming that the conductor is a straight round wire, the inductance can be determined from (see Section 5.2.2.3):

$$L = 0.002 \ell \left( 2.303 \log \frac{4\ell}{d} - 0.75 \right), \quad (3-8)$$

where  $L$  is the total inductance in microhenries,  $\ell$  is the length in cm, and  $d$  is the diameter in cm. A 30-meter length of conductor will exhibit an inductance of 52.5 microhenries.

The voltage,  $V$ , developed across an inductance is given by

$$V = L \, di/dt, \quad (3-9)$$

where  $L$  is the inductance in henries and  $di/dt$  is the rate of change of the current through the inductor in amperes per second. From Table 3-1, the rate of rise of the typical lightning stroke is 20 kA/ $\mu$ s which corresponds to a  $di/dt$  of  $2 \times 10^{10}$  amps/second. Thus the voltage developed by the discharge pulse through the 30-meter (100 foot) downconductor is

$$V = 5.25 \times 10^{-5} \times 2 \times 10^{10} = 1.05 \times 10^6 \text{ volts.} \quad (3-10)$$

Although the duration of this voltage is typically less than 2 microseconds, the voltage generated is high enough to cause flashover to conducting objects located as much as 35 cm (14 in.) away from the down conductor. It is for this reason that metallic objects within 6 feet of lightning down conductors should be electrically bonded to the down conductors.

### 3.6.3.2 Induced Voltage Effects.

In addition to the lightning effects discussed above, circuits not in direct contact with the lightning discharge path can experience damages even in the absence of overt coupling by flashover. Because the high current associated with a discharge exhibits a high rate of change, voltages are electromagnetically induced on nearby conductors. Experimental and analytical evidence (3-12) shows that the surges thus induced can easily exceed the tolerance level of many components, particularly solid state devices. Surges can be induced by lightning current flowing in a down conductor or structural member, by a stroke to earth in the vicinity of buried cables, or by cloud-to-cloud discharges occurring parallel to long cable runs, either above ground or buried (3-2).

Consider a single-turn loop parallel to a lightning down conductor such as that shown in Figure 3-9. The voltage  $E$  magnetically induced in the loop is related to the rate of change of flux produced by the changing current in the down conductor (see Section 6.2.2.1). The voltage induced in the loop is dependent upon the dimensions of the loop ( $l$ ,  $r_2 - r_1$ ), its distance from the down conductor ( $r_1$ ), and the time rate of change of the discharge current ( $di/dt$ ). Figure 3-10 is a plot of normalized voltage per unit length that would be developed in a single turn loop of various widths.

These results suggest the steps that should be taken to minimize the voltage induced in signal, control, and power lines by lightning discharges through down conductors. First, since no control can be exercised over  $di/dt$  because it is determined by the discharge itself,  $E$  must be reduced by controlling  $l$ ,  $r_1$  and  $r_2$ . The variable  $l$ , is a measure of the distance that the loop runs parallel to the discharge path; thus, by restricting  $l$ , the induced  $E$  can be minimized. Thus cables terminating in devices or equipments potentially susceptible to voltage surges should not be run parallel to conductors carrying lightning discharge currents if at all possible. If parallel runs are unavoidable, Figure 3-10 also shows that the distance,  $r_1$ , between the loop and the lightning current path should be made as large as possible.

Another observation to be made from Figure 3-10 is that  $r_2$  minus  $r_1$  should be as close as possible to zero. In other words, the distance between the conductors of the pickup loop should be minimized. One common way of reducing this distance is to twist the two conductors together such that the average distance from each conductor to the discharge conductor is the same.

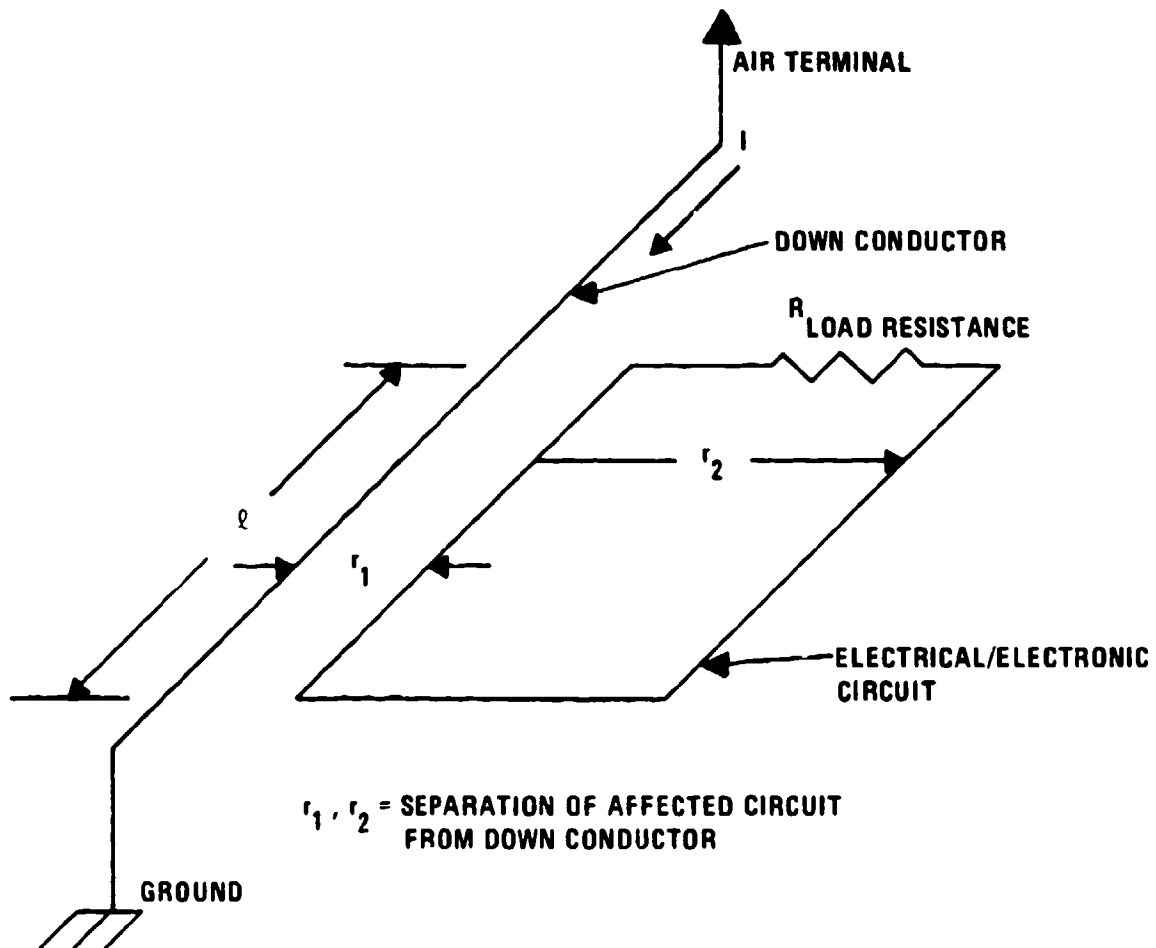


Figure 3-9. Inductive Coupling of Lightning Energy to Nearby Circuits

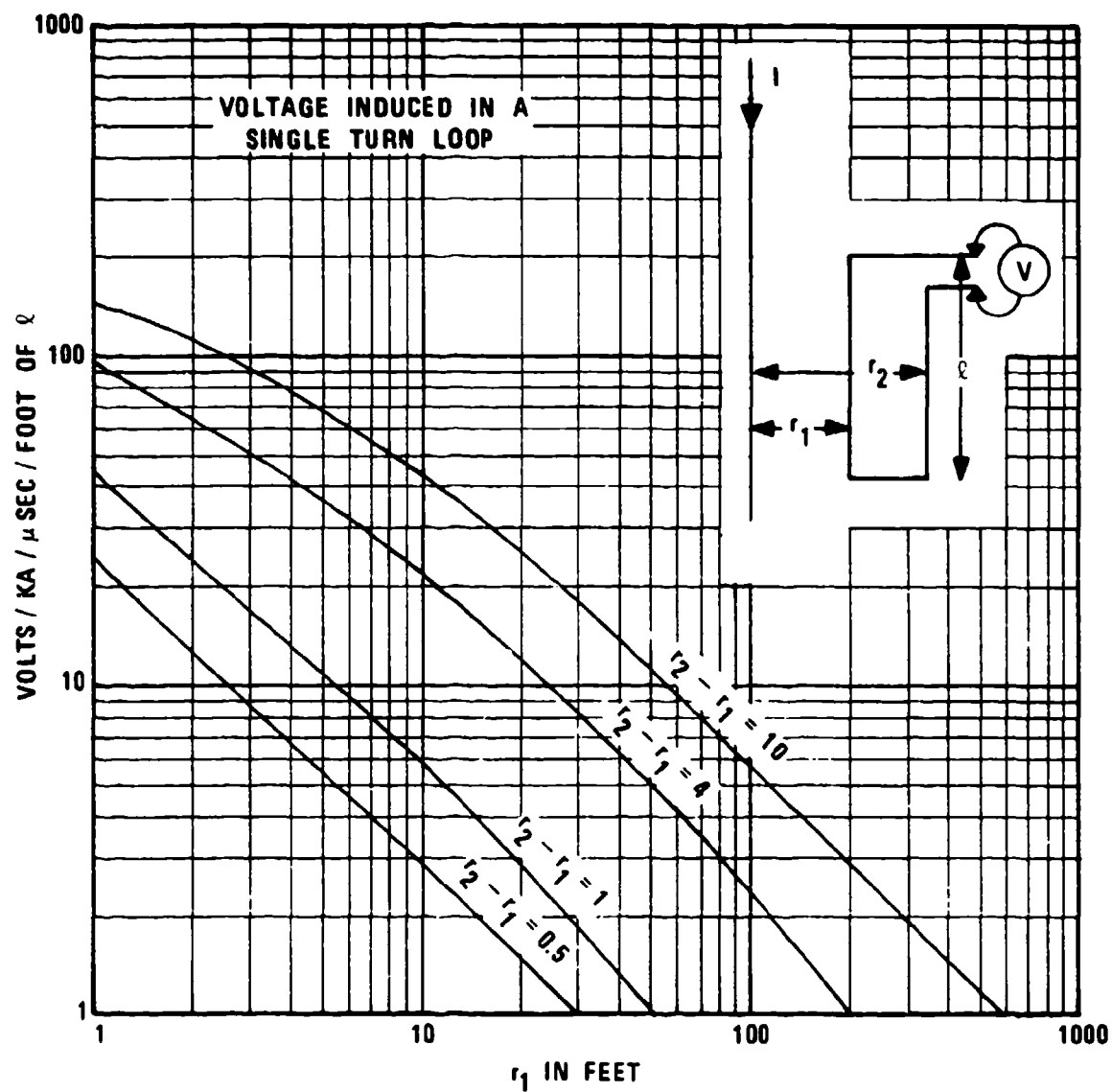


Figure 3-10. Normalized Voltage Induced in a Single-Turn Loop by Lightning Currents

Another protective measure is to reduce the flux density within the pickup loop by providing magnetic shielding. Because the coupling field is primarily magnetic in nature, a shielding material having a high permeability such as iron or nickel should be used. For shielding against lightning-produced fields, steel conduit or cast iron pipe are much more effective than aluminum or other non-ferrous materials.

### 3.6.3.3 Capacitively-Coupled Voltage.

Prior to the lightning discharge, an electric charge slowly accumulates on earth-based objects in the vicinity of the electrified clouds. This increase in charge occurs slowly enough so that the potential of grounded conductors does not change appreciably with respect to the earth, even when the impedance to ground is high. When the lightning stroke terminates on a structure or other point having contact with the earth as illustrated in Figure 3-11, the charge on all grounded objects nearby suddenly becomes redistributed. The redistribution of charge produces a current flow through the grounding impedance of the grounded objects and produces a voltage across that impedance.

Referring to Figure 3-11, the voltage between the conducting objects and the ground can be expressed as

$$E = \frac{Q}{C} e^{-t/RC} \quad (3-11)$$

where  $Q$  is the stored charge in coulombs,  $C$  is the total capacitance to ground in farads,  $R$  is the effective resistance to ground in ohms, and  $t$  is the elapsed time in seconds from the occurrence of the stroke.

Equation 3-11 shows that if the product  $RC$  is small, the exponential term will be large (for a time  $t$  on the order of 10  $\mu$ s), thus making the voltage capacitively induced on any reasonably well-grounded object quite small for a typical lightning stroke.

### 3.6.3.4 Earth Resistance.

Consider a facility such as the one illustrated in Figure 3-12, that has more than one possible electrical path to earth. For example, a ground rod is driven into the earth at the transformer pole or at the service entrance to Building 1. The resistance,  $R_{G1}$ , of this rod could be 25 ohms or higher and still conform to NEC requirements. Metal utility pipes such as water lines generally offer a relatively low resistance (labeled  $R_{G2}$ ) to earth. (In soils of high resistivity the point of effective contact between utility pipes may be an appreciable distance from the facility.) Empirical data indicates that the grounding resistance offered by water pipes is on the order of 1 to 3 ohms. If the electrical ground is not connected to the water pipe, a lightning strike to the ground wire of the electrical distribution system could produce a potential difference high enough to possibly produce an arc between the electrical ground (including the equipment cabinet and the building's structure, if connected) and the utility piping. A definite personnel hazard would then exist because of the high voltage that would be developed between the equipment and building ground and pipes. Because of this reason as well as the requirement to prevent analogous hazards from existing during power system faults, MIL-STD-188-124A requires electrical safety grounds be connected to the metallic water system in the building and recommends they also be directly connected to the ground rod at the transformer.

If  $R_{G1}$  is 25 ohms while  $R_{G2}$  is only 1 ohm or so, then a lightning strike as indicated could easily cause the potential of the overhead ground wire to become high enough to produce an arc across the transformer windings and insulators. Since the low voltage secondary side offers a lower impedance to earth, it is the preferred path for the discharge.

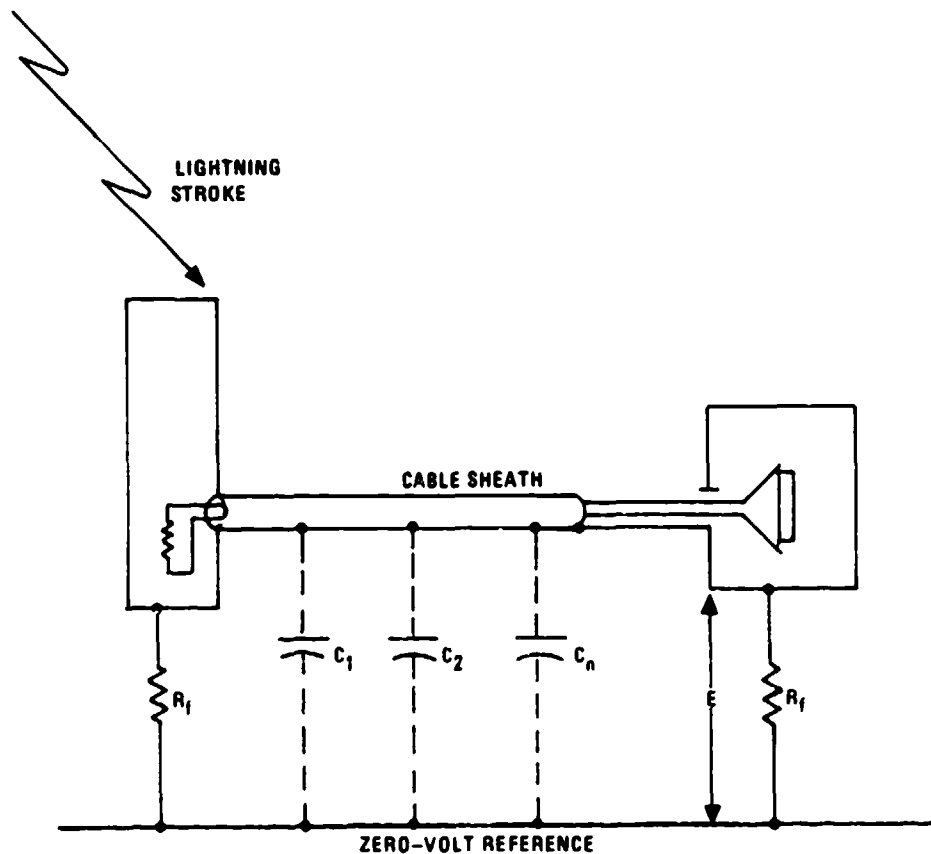


Figure 3-11. Capacitive Coupling of Lightning Energy

This type of lightning threat can be minimized by (1) reducing  $R_{G1}$  to approximately the magnitude of  $R_{G2}$ , (2) the installation of appropriate lightning arresters at the transformer to keep the potential difference between the power conductors and the ground wire and between the primary and secondary windings to within the stress ratings of the transformer, and (3) interconnecting the earth electrode subsystem (to include the water and other utility pipes) with a 1/0 or larger buried copper cable as illustrated by the dotted line in Figure 3-12.

Interconnecting the ground electrodes of the building and transformer pole to form one effective earth contact does not eliminate the lightning threat to the buried cable between the two buildings. As shown, the cable shield is connected to the cabinet, i.e., the building ground. In the event of a lightning strike as shown, Building 1 and its power supply system will be elevated in potential relative to Building 2. In particular, if the distance between the two buildings is more than just a few meters, the inductance, primarily, of the cable shield will prevent the cable from providing the low impedance necessary to keep the two buildings at the same potential. In addition if the shield of the cable is insulated from the earth, as is usually the case, the potential of the cable shield can become high enough with respect to the earth to exceed the breakdown of the insulation.

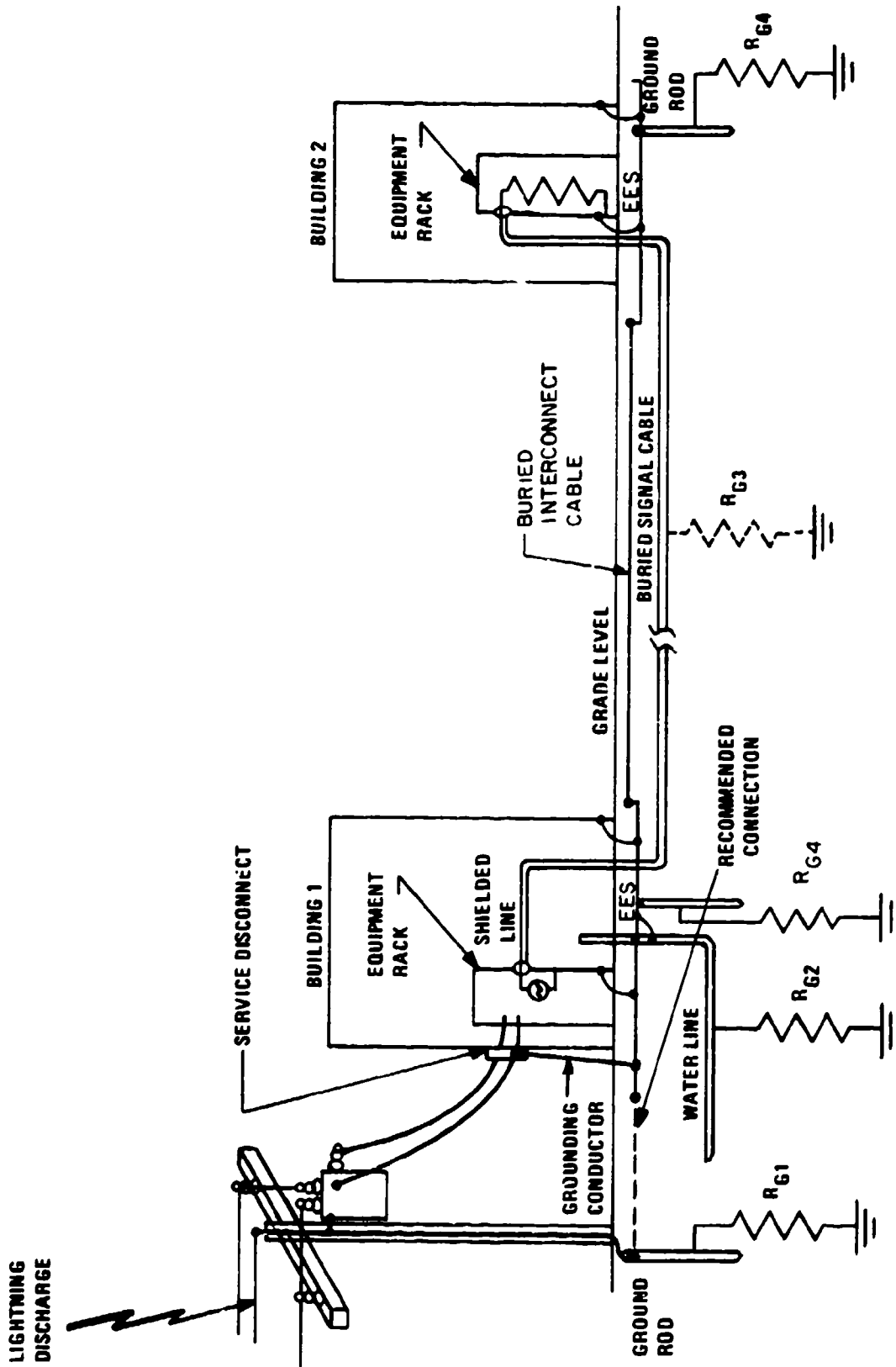


Figure 3-12. Coupling of Lightning Energy Through and Interconnected Facility



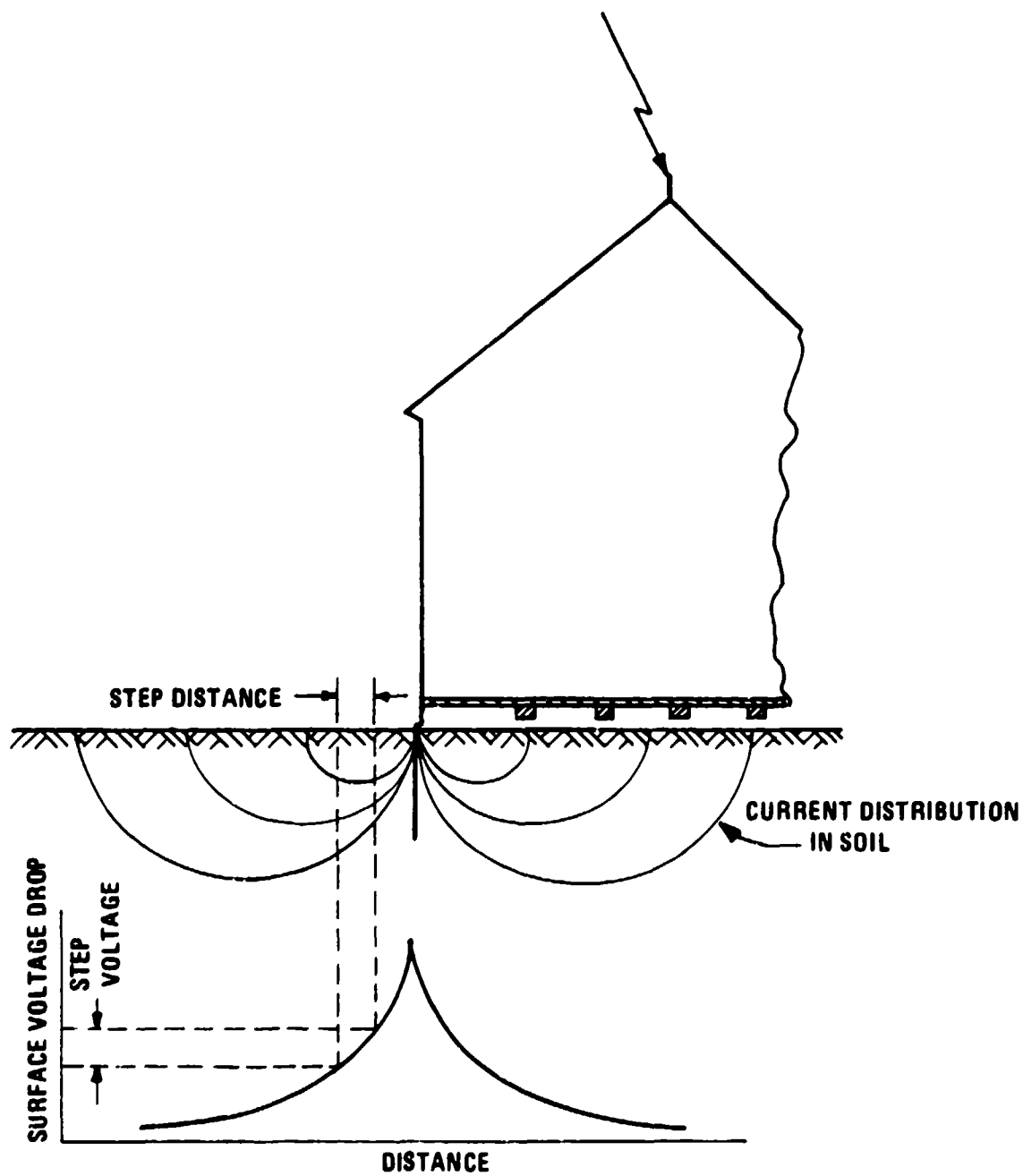


Figure 3-13. Step-Voltage Hazards Caused by Lightning-Induced Voltage Gradients in the Earth

Assume for the moment that Building 1 has an earth electrode subsystem consisting of ground rods interconnected with the cold water system with a net resistance to earth of 3 ohms. With a lightning discharge of 20 kA, the voltage of the complex will rise to 60 kV with respect to Building 2 and that portion of the earth not in the immediate vicinity of Building 1. At Building 1, the cable shield voltage will rise along with that of the building. This voltage pulse will travel down the cable, successively raising the shield potential to as much as 60 kV with respect to the surrounding earth. Such high voltages cause insulation breakdown in the form of tiny pinholes where the lightning energy punches through.

As the lightning pulse travels down the cable, its amplitude diminishes due to cable resistance and dielectric losses. However, the amplitude of the pulse can still be sufficient to damage circuit components in terminating equipment in Building 2. To minimize this damage, surge arresters compatible with the terminating components and hardware should always be provided on such cables. Further information on the use of surge arresters is presented in Volume II, Section 1.3.3.5.

In the event of a lightning stroke, there is a definite personnel hazard posed by the voltage gradient in the soil in the vicinity of the point where the lightning discharge enters the earth. In homogeneous soil, the current rapidly leaves the electrode. The current density is highest near the electrode and rapidly decreases with distance from the electrode. In soil of uniform resistivity, a significant voltage gradient will exist between two points that are differing distances from the electrode. Figure 3-13 illustrates the nature of this voltage variation and shows the hazard encountered by personnel walking (or standing) in the area. The voltage difference across the span of a step can be sufficient to be lethal. As shown earlier, the degree of the hazard is determined by the magnitude of the stroke current, the grounding resistance of the earth electrode, and the distance away from the electrode. No control can be exercised over the current; the threat, however, can be lessened by achieving a low common ground resistance and by minimizing the step potential as discussed in Section 2.8.1.3.

### 3.7 BASIC PROTECTION REQUIREMENTS.

To effectively protect a structure such as a building, mast, tower, or similar self-supporting object from lightning damage, the following requirements must be met:

a. An air terminal of adequate height, mechanical strength and electrical conductivity to withstand the stroke impingement must be provided to intercept the discharge to keep it from penetrating any nonconductive outer coverings of the structure or to prevent it from terminating on antennas, lighting fixtures, transformers, or other devices likely to be damaged or destroyed.

b. A low impedance path from the air terminal to earth must be provided.

c. The resistance of the connection between the discharge path and the earth must be low.

These requirements are met by either (1) an integral system of air terminals, roof conductors, and down conductors, securely interconnected to provide the shortest practicable path to earth, or (2) a separately mounted shielding system such as a metal mast which acts as an air terminal, and a down conductor or an overhead ground wire terminated at the ends (and at intermediate locations, if needed) with down leads connected to earth ground electrodes. Specific design practices are contained in Volume II.

### 3.8 DETERMINING THE NEED FOR PROTECTION.

The degree to which lightning protection is required, is a subjective decision requiring an examination of the relative criticalness of the structure location and its contents to the overall mission of the facility. Those structures containing elements vital to the operational mission such as air traffic control towers, radar installations, navigational aids, and communications centers are examples of facilities which obviously must be protected. However, every building or structure does not require that a lightning protection system be installed. For example, buildings primarily used for the storage of nonflammable materials do not have a critical need for protection.

Three of the factors to consider in ascertaining whether a given structure should have a lightning protection system installed or in determining the relative comprehensiveness of the system are the relative threat of being struck by lightning, the type of construction, and the nature of the facility.

#### 3.8.1 Strike Likelihood.

The relative likelihood of a particular structure being struck by lightning is a function of the keraunic level, i.e., the thunderstorm activity of the locality, the effective height of the structure and its attractive area. Average thunderstorm activity can be determined from the isokeraunic maps shown in Figures 3-2 and 3-3. Then using the techniques described in Section 3.4, estimate the frequency with which strikes to the structure may occur. Use this estimation as one of the inputs to the decision process.

#### 3.8.2 Type of Construction.

Steel frame buildings with metal outer coverings offer the greatest inherent protection against lightning damage. Steel towers also exhibit a high immunity to structural damage. Additional protection for these type buildings will probably be required only for very critical facilities in highly exposed locations. Steel frame buildings with nonconductive, but nonflammable, outer coverings (like brick or other masonry) also offer a high degree of protection against lightning damage. The greatest hazard is posed by pieces of masonry being dislodged by stroke currents passing through the outer coverings to reach the structural steel underneath. Minimal protection consisting of interconnected air terminals to down conductors and steel support columns will be sufficient to prevent this type of structural damage.

Buildings constructed of nonconductive materials such as wood, concrete blocks, or synthetic materials are the most susceptible to destructive damage. A complete auxiliary protection system will be required to prevent lightning damage to buildings utilizing this type of construction.

### 3.8.3 Criticalness to System Mission.

If a strike to the facility poses a threat to human life, either to the occupants of the structure or to those persons whose safety is dependent upon reliable performance of the equipment and people inside the structure, comprehensive lightning protection should be definitely provided even in areas of low thunderstorm activity. At the other extreme, the need for the protection of buildings used primarily to store nonflammable or nonexplosive items is doubtful unless the stored items are critical to system operation, the building is usually exposed, etc. In between these extremes are those structures whose incapacitation would cause an inconvenience or present other difficulties short of life-and-death situations. With these structures, a careful analysis must be made to determine the relative likelihood of outages from lightning in comparison to normal equipment failures, downtime for maintenance, and other routine occurrences.

Though not directly related to the protection of electrical or electronic installations, Reference 3-10 is recommended for further guidance in performing the tradeoff analyses to determine the degree of lightning protection required for specific facilities.

### 3.9 APPLICABLE CODES.

The Lightning Protection Code, NFPA No. 78, issued by the National Fire Protection Association (3-9) contains the basic requirements for the minimization of personnel hazards in the event of a lightning strike to the structure.

The requirements of NFPA No. 78, however, are not sufficient to protect the electrical distribution system, signal and control cables, or sensitive electronic equipment from surges produced by either direct or indirect strokes. Thus additional steps such as providing lightning arresters on power lines and on outside signal and control cables, providing counterpoise cables for overhead and underground cables, providing comprehensive electromagnetic shielding on sensitive cables, and installing fast response surge protection devices on circuits exposed to lightning discharges should be taken. MIL-STD-188-124A refers.

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## CHAPTER 4

## FAULT PROTECTION SUBSYSTEM

**4.1 FAULT PROTECTION.** For effective fault protection, a low resistance path must be provided between the location of the fault and the transformer supplying the faulted line. The resistance of the path must be low enough to cause ample fault current to flow and rapidly trip breakers or blow fuses. The necessary low resistance return path inside a building is provided by the grounding (green wire) conductor and the interconnected facility ground network. An inadvertent contact between energized conductors and any conducting object connected to the grounding (green wire) conductor will immediately trip breakers or blow fuses. In a building containing a properly installed third-wire grounding network, as prescribed by MIL-STD-188-124A, faults internal to the building are rapidly cleared regardless of the resistance of the earth connection.

**4.1.1 Power System Faults.**

A power system fault is either a direct short or an arc (continuous or intermittent) in a power distribution system or its associated electrical equipment. These faults are hazardous to personnel for several reasons:

- a. Fault currents flowing in the ground system may cause the chassis of grounded equipment to be at a hazardous potential above ground.
- b. The energy in a fault arc can be sufficient to vaporize copper, aluminum, or steel. The heat can present a severe burn hazard to personnel.
- c. There is a fire hazard associated with any short circuit or arc.
- d. Burning insulation can be particularly hazardous because of the extremely toxic vapors and smoke which may be produced.

Some common causes of electrical system faults are:

- a. Rodents getting between ground and phase conductors.
- b. Water infiltration.
- c. Moisture in combination with dirt on insulator surfaces.
- d. Breakdown of insulation caused by thermal cycling produced by overloads.
- e. Environmental contaminants.
- f. Damage during installation.
- g. System age deterioration.

Figure 4-1 illustrates how personnel hazards are developed by improper installation and fault conditions. Suppose that one phase of the 230-volt line accidentally contacts the motor frame. If the motor is not grounded, its frame will rise to 133 volts, and anyone coming in contact with it would be subject to a lethal shock if simultaneous contact is made with a grounded object. To prevent this situation from arising, the motor frame must be grounded via the green wire. The resistance of the fault path must be low enough to permit the fault current to trip the overload protector and interrupt the fault. If the resistance of the fault path is too large, the fault current will not be enough to trip the overload protectors. Thus to minimize both shock and fire hazards, the resistance of the fault path must be as low as possible. However, the fault protection subsystem normally does not depend on the earth electrode subsystem to trip overcurrent devices. The fault current normally flows through the green wire (grounding conductor) to the source side of the first service disconnect means where the green wire and the neutral are tied together. The fault current then flows through the neutral to the transformer to complete the circuit. This path functions completely independent of the connection to the earth electrode subsystem.

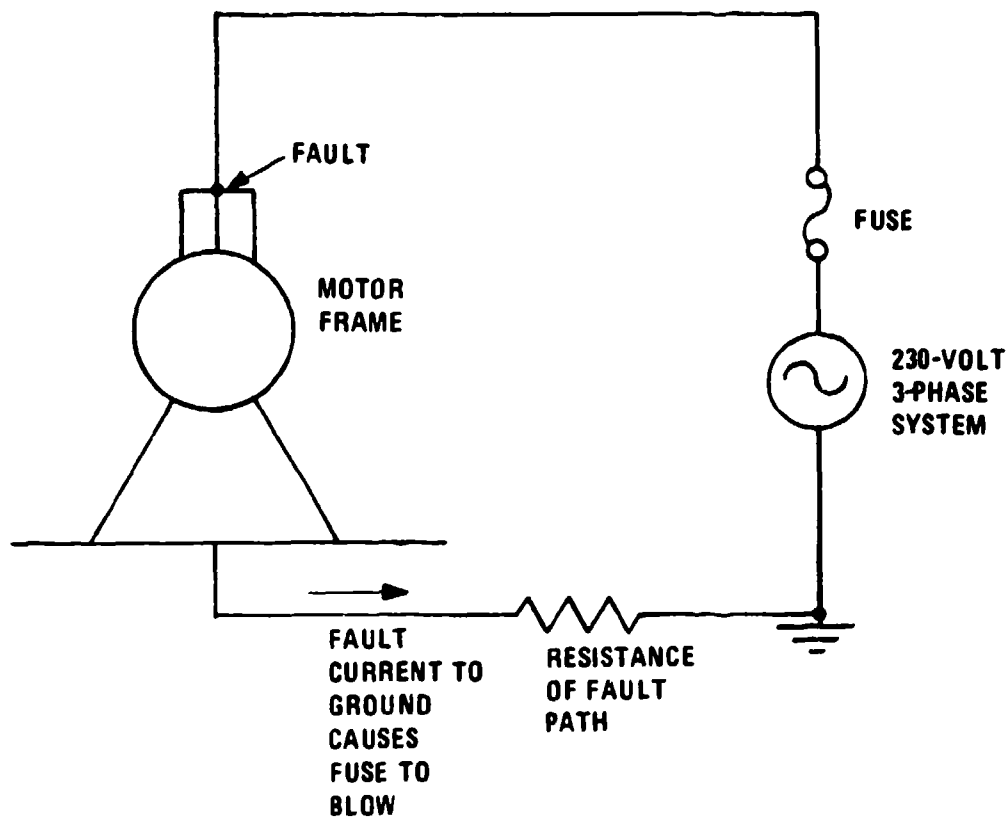


Figure 4-1. Grounding for Fault Protection

Fault clearance in power distribution systems is normally provided by circuit breakers, fuses, or overload relays in each phase. These devices provide personnel protection only if the fault current is sufficient to trip the over-current device. They generally however do not have response times which are adequate to protect the individual if he happens to be in direct contact with the energized object.

**4.1.2 Ground-Fault-Circuit-Interrupter (GFCI).** High resistance faults (low and moderate currents of 5 milliamperes or more) can be cleared rapidly with a device called a ground-fault-circuit-interrupter (GFCI). The GFCI contains an electronic circuit which continuously monitors the difference between the current supplied to the load and the current returned from the load. If this difference is not zero, some current must be leaking to ground. When this leakage current exceeds a preset value, the GFCI will act to interrupt the power to the circuit. GFCIs are so sensitive that they can be set to interrupt power fault currents as low as 2 milliamperes. Experiments with dogs have shown that trip currents of 5 milliamperes or less will prevent electrocution. (GFCIs have proven so effective as protection against electric shock that the National Electrical Code requires that all 15 and 20 ampere bathroom, garage, and outdoor receptacles in family dwelling units and in circuits set up at construction sites be protected with a GFCI. MIL-STD-188-124A also recommends they be installed on 120 volt single phase 15 and 20 ampere receptacles of C-E facilities.)

## **4.2 EARTH CONNECTION.**

Historically, grounding requirements arose from the need to protect personnel, equipment, and facilities from lightning strokes and from industrially generated static electricity. Structures, as well as electrical equipment, were connected to earth, i.e., grounded, to provide the path necessary for lightning and static discharges. As utility power systems developed, grounding to earth was found to be necessary for safety. All major components of the system such as generating stations, substations, and distribution systems are earth grounded to provide a path back to the generator for the fault currents in case of transmission line trouble. The path to earth should have as low a resistance as possible. A low resistance minimizes the potential difference between equipments connected to the earth electrode subsystem when fault currents flow. Thus personnel who come in contact with two or more pieces of equipment at one time are protected.

Ideally, the earth connection should exhibit zero resistance between the earth and the equipment and facilities connected to it. Any physically realizable connection, however, will exhibit a finite resistance to earth. The economics of the design of the earth electrode subsystem involves a trade-off between the expense necessary to achieve a low resistance and the satisfaction of minimum subsystem requirements. The 10 ohm design objective of MIL-STD-188-124A is considered such a trade-off.

## **4.3 AC POWER LINE GROUND.**

The grounding conductor (green wire) in a single-phase 115/230 volt ac power distribution system in a facility is one of four leads, the other three being the two phase or "hot" leads (black/red) and the neutral lead (white wire). The green wire is a safety conductor designed to carry current only in the event of a fault. The "hot" leads are connected from the first service disconnect to the high sides of the secondary of the distribution transformer and the neutral is connected to the center tap which is grounded to a ground terminal at the transformer. When a single transformer supplies power to any one communications building, for fault



protection the grounding conductor shall be grounded on the source side of the first service disconnect to the earth electrode subsystem and also to the ground terminal at the distribution transformer. For 3-phase wye systems a five-wire service entry cable consisting of one neutral, one grounding, and three phase conductors shall be employed. In either case, when a single transformer supplies power to a single building, the safety ground (green wire) shall be grounded to the earth electrode subsystem at the supply side of the first service disconnect of the facility as well as at the distribution transformer as shown in Figure 4-2. The neutral shall also be grounded at both locations.

When a single transformer supplies power to more than one C-E building and if noise or hum is encountered in C-E circuits or equipments, the neutral should be lifted or removed from ground at each service disconnect. In this case the neutrals from each building are grounded at the distribution transformer only (see Figure 4-3).

To protect personnel from exposure to hazardous voltages, all exposed metallic elements of electrical and electronic equipment are connected to ground with the green wire. Then, in the event of inadvertent contact between the hot lead and chassis, frame, or cabinet through human error, insulation failure, or component failure, a direct fault clearance path is established to quickly remove the hazard.

Grounding of a 3-phase wye power distribution system is accomplished similarly to the single phase system. The connections for a typical system are shown in Figure 4-3. As in single phase systems, the neutral lead is bonded to the green wire at the supply side of the first service disconnecting means and grounded to the earth electrode subsystem as well as to the ground terminal at the distribution transformer. If one transformer supplies power to more than one C-E building, the neutral is lifted from ground at the service disconnect.

A 3-phase system served by a transformer with a delta connected secondary will require the use of a grounding transformer to ground the system and establish a neutral. The grounding transformer may be either a "zig-zag" or "wye-delta" type, both of which have leads which are attached to each of three phases and a fourth lead which is grounded and serves as the neutral. The typical connections for a grounding transformer are shown in Figure 4-4.

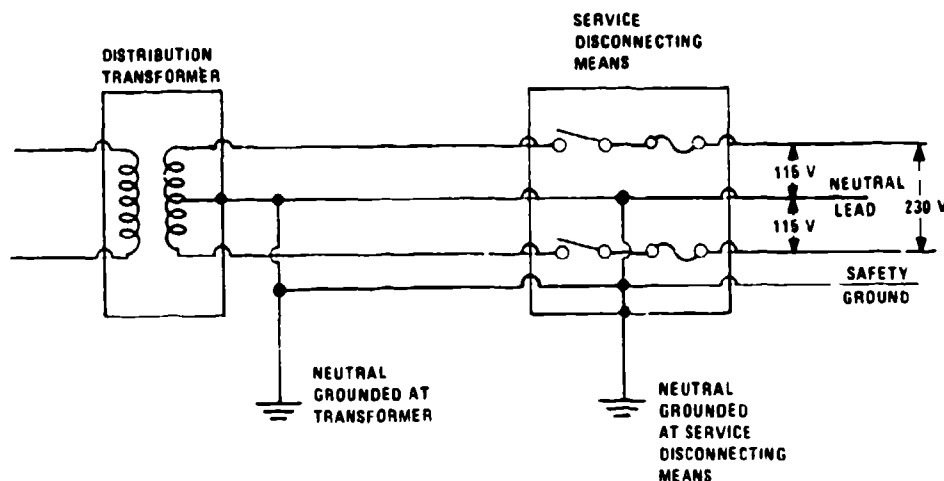
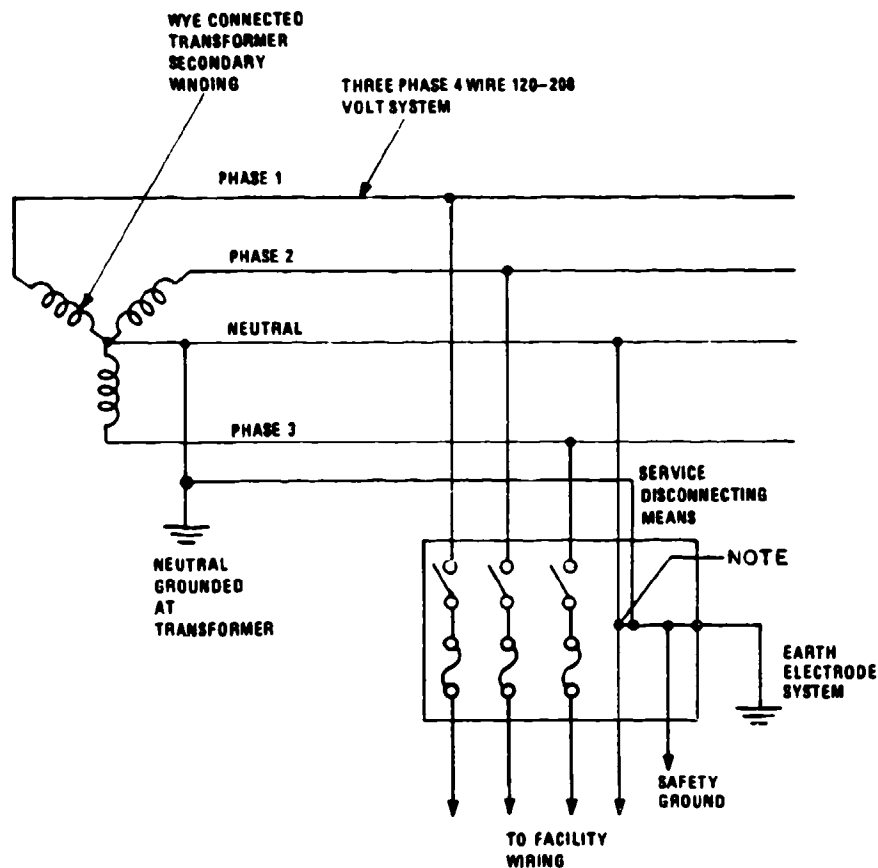


Figure 4-2. Single-Phase 115/230 Volt AC Power Ground Connections

**4.4 TEST EQUIPMENT.** Test equipments are available to measure the resistances and impedances of the fault protection subsystems including the grounding (green) conductor as well as the signal reference subsystem (equipotential plane) which may at times become part of the fault protection subsystem. These equipments can measure the impedances (at 60 Hz) of each path from the equipment having the fault to the first service disconnect means and therefore assist in determining the value of the fault current over each path. The information will in turn be beneficial in determining or predicting the degree of interference which may be anticipated should a fault current be superimposed on the signal reference subsystem. (4-1 and 4-2)



NOTE: Lift when single transformer supplies power to more than one building or because of objectionable current, noise or interference.

Figure 4-3. Three-Phase 120/208 Volt AC Power System Ground Connections

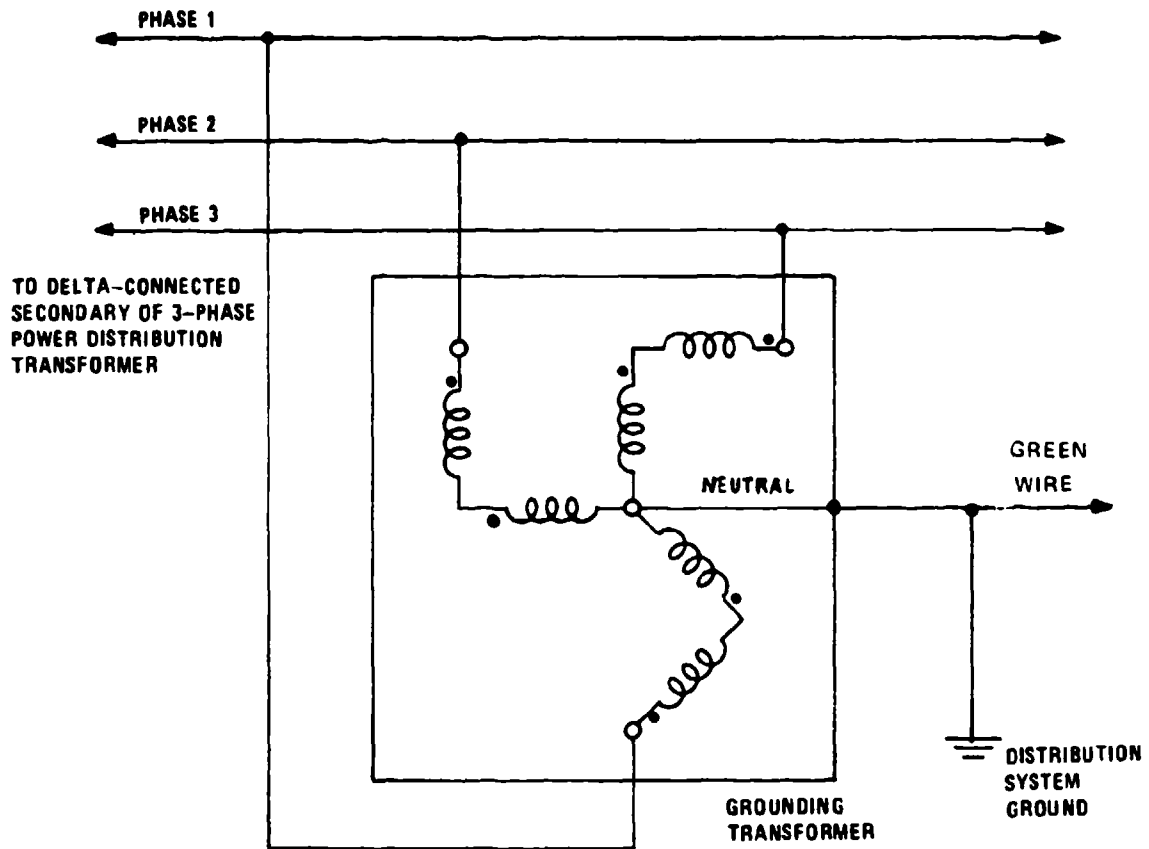


Figure 4-4. Connections for a Three-Phase "Zig-Zag" Grounding Transformer

#### 4.5 REFERENCES.

- 4-1. A Practical Approach to Establish Effective Grounding for Personnel Protection, IEEE Conference Paper, Chris C. Kleronomos and Edward Cantwell, 1979.
- 4-2. Some Fundamentals of Equipment Grounding Circuit Design, R.H. Kaufmann, 1954.

## CHAPTER 5

## GROUNDING OF SIGNAL REFERENCE SUBSYSTEM

5.1 INTRODUCTION.

Signal circuits are grounded and referenced to ground to (1) establish signal return paths between a source and a load, (2) control static charge, or (3) provide fault protection. The desired goal is to accomplish each of these three grounding functions in a manner that minimizes interference and noise.

If a truly zero impedance ground reference plane or bus could be realized, it could be utilized as the return path for all currents -- power, control, audio and rf -- present within a system or complex. This ground reference would simultaneously provide the necessary fault protection, static discharge, and signal returns. The closest approximation to this ideal ground would be an extremely large sheet of a good conductor such as copper, aluminum, or silver underlying the entire facility with large risers extending up to individual equipments. The impedance of this network at the frequency of the signal being referenced is a function of conductor length, resistance, inductance, and capacitance. When designing a ground system in which rf must be considered, transmission line theory must be utilized.

5.2 CONDUCTOR CONSIDERATIONS.5.2.1 Direct Current Resistance.

The resistance,  $R_{dc}$ , of a conductor of uniform cross section is proportional to the length and inversely proportional to the cross-sectional area, that is

$$R_{dc} = \rho l / A \text{ ohms ,} \quad (5-1)$$

where  $\rho$  is the resistivity of the conductor material,  $l$  is the length of the conductor in the direction of current flow, and  $A$  is the cross-sectional area of the conductor. Values of  $R_{dc}$  for the standard sizes of wire and cable are given in Table 5-1. (For data on wire sizes not shown in this table, consult References 5-1 and 5-2.)

At dc, the resistance of the conductor is the controlling factor. Except for very unusual situations (such as when the signal to be processed is very low in amplitude or where the interfacing equipments are very far apart physically), an adequate ground can generally be realized for dc in a relatively economical manner utilizing low resistivity materials such as copper and aluminum. Most systems, however, employ other than dc signals. Therefore, the frequency-dependent properties of the conductors become important.

5.2.2 Alternating Current Impedance. The ac impedance of a conductor is composed of two parts: the ac resistance and the reactance. Both the ac resistance and the reactance of a conductor vary with frequency as a result of skin effect.

Table 5-1

## Properties of Annealed Copper Wire

AWG No.	Diameter		Cross-Sectional Area		Resistance in Ohms	
	<u>mils</u>	<u>mm</u>	<u>cmil</u>	<u>mm<sup>2</sup></u>	<u>per 1000 ft</u>	<u>per km</u>
4/0	460.0	11.7	211 600	107.2	0.049	0.161
3/0	409.6	10.4	167 800	85.0	0.062	0.203
2/0	364.8	9.3	133 100	67.4	0.078	0.256
1/0	324.9	8.3	105 500	53.4	0.098	0.322
1	289.3	7.3	83 690	42.4	0.124	0.407
2	257.6	6.5	66 370	33.6	0.156	0.512
4	204.3	5.2	41 740	21.1	0.248	0.814
6	162.0	4.1	26 250	13.3	0.395	1.296
8	128.5	3.3	16 510	8.4	0.628	2.060
10	101.9	2.6	10 380	5.3	0.999	3.278
12	80.8	2.1	6 530	3.3	1.588	5.210
14	64.1	1.6	4 107	2.1	2.525	8.284
16	50.8	1.3	2 583	1.3	4.016	13.176
18	40.3	1.0	1 624	0.8	6.385	20.948
20	31.9	0.8	1 022	0.5	10.150	33.300

5.2.2.1 Skin Effect.

Whereas a direct current is uniformly distributed over the cross-sectional area of a conductor, alternating current tends to concentrate near the surface of the conductor. The higher the frequency, the greater the concentration near the surface. This physical phenomenon is called skin effect. A measure of the degree of penetration of the currents into the conductor is given by the skin depth,  $\delta$ .  $\delta$  is defined as the depth at which the current density is attenuated to  $1/e = 1/2.718 \approx 0.37$  of its value at the conductor surface. Skin depth may also be interpreted as the equivalent thickness of a hollow conductor carrying a uniform distribution over its cross-sectional area, having the same external shape as the actual conductor, and having a dc resistance exactly the same as the ac resistance of the conductor.

For conductors whose thickness is at least three times the skin depth, this depth is given by (5-3).

$$\delta = 5033 \sqrt{\frac{\rho}{\mu_r f}} \text{ cm}, \quad (5-2)$$

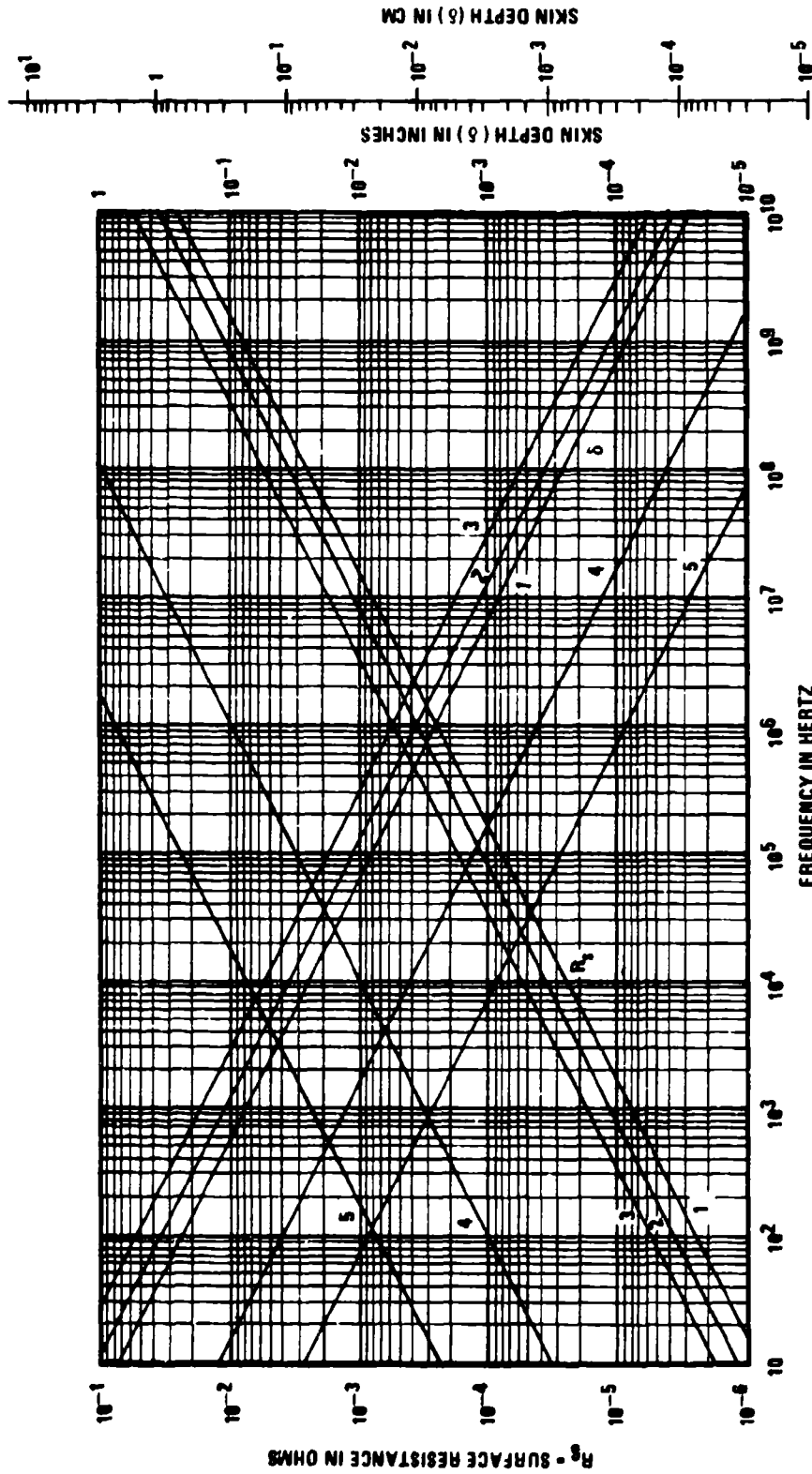
where  $\rho$  is the resistivity of the material in ohm-cm,  $f$  is the frequency in hertz, and  $\mu_r$  is the relative permeability of the material. The skin depth for various metals is given in Table 5-2 and Figure 5-1. Note that copper has a skin depth of 0.34 inch (8.63 mm) at 60 Hz but only .00026 inch (0.066 mm) at 1 MHz.

Table 5-2

Parameters of Conductor Materials (5-4)

Material*	$\frac{\rho}{(\Omega\text{-cm})}$	$\frac{\delta}{(\text{cm})}$	$\frac{R_s}{(\Omega)}$
Silver	$1.62 \times 10^{-6}$	$6.41/\sqrt{f}$	$2.52 \times 10^{-7} \sqrt{f}$
Copper	$1.73 \times 10^{-6}$	$6.62/\sqrt{f}$	$2.61 \times 10^{-7} \sqrt{f}$
Aluminum	$2.69 \times 10^{-6}$	$8.25/\sqrt{f}$	$3.26 \times 10^{-7} \sqrt{f}$
Brass	$6.37 \times 10^{-6}$	$12.70/\sqrt{f}$	$5.01 \times 10^{-7} \sqrt{f}$
Solder	$14.2 \times 10^{-6}$	$18.96/\sqrt{f}$	$7.48 \times 10^{-7} \sqrt{f}$

\*  $\mu_r = 1$



- (1) COPPER  
(2) ALUMINUM  
(3) BRASS  
(4) SOFT IRON  
(5) PERMALLOY 70

Figure 5-1. Surface Resistance and Skin Depth for Common Metals (5-5)

### 5.2.2.2 AC Resistance.

The ac resistance of a conductor of any shape can be determined from the skin depth if both the thickness and the radius of curvature of the conductor are much greater than the skin depth and if the radius of curvature does not vary too rapidly around the conductor's perimeter. For a conductor meeting these conditions, the ac resistance per unit length is

$$R_{ac} = \frac{c}{\delta P} \quad \text{ohms/meter, or} \quad (5-3)$$

$$= \frac{R_s}{P} \quad \text{ohms/meter} \quad (5-4)$$

where  $P$  is the circumference of the conductor and  $R_s$  is the surface resistance of the conductor. The surface resistance is defined as the ac resistance of a surface of equal length and width and is given by

$$R_s = \frac{\rho}{\delta} = 1.987 \times 10^{-4} \sqrt{\mu_r f \rho} \quad (5-5)$$

The surface resistance for various metals is also shown in Figure 5-1 and Table 5-2.

The ratio of the ac resistance to the dc resistance is called the resistance ratio of a conductor. Skin effect causes the resistance ratio to be greater than unity. The resistance ratio for straight cylindrical wires is given in Figure 5-2 in terms of a parameter  $X$  defined as

$$X = \sqrt{\frac{8\pi\mu_r f}{R_{dc} \times 10^9}} \quad (5-6)$$

where  $\mu_r$  is the relative permeability of the conductor,  $f$  is the frequency in hertz, and  $R_{dc}$  is the dc resistance in ohms for 1 cm of conductor.\* In the case of copper wire, Equation 5-6 becomes

$$X_{cu} = 2.71 \times 10^{-4} d_m \sqrt{f} \quad (5-7)$$

where  $d_m$  is the wire diameter in mils, or becomes

$$X_{cu} = 1.07 \times 10^{-2} d_m \sqrt{f} \quad (5-8)$$

where  $d_m$  is diameter in mm.

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\*It should be noted that Equation 5-6 applies at all frequencies, whereas Equations 5-3 and 5-4 apply only under the conditions stated.



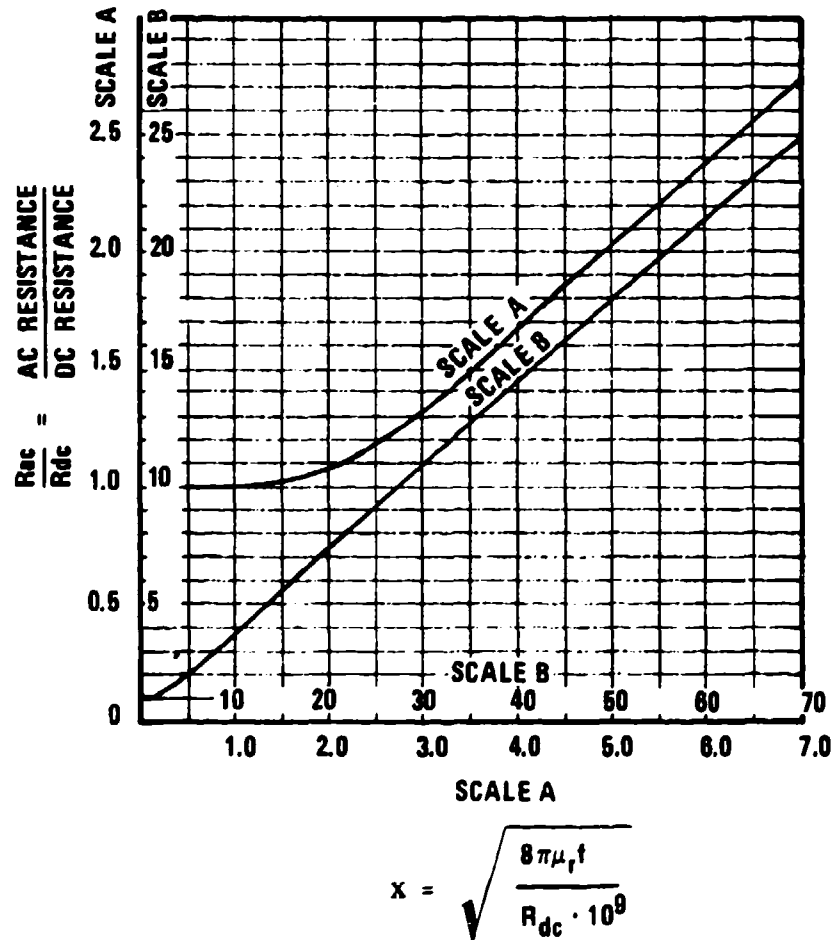


Figure 5-2. Resistance Ratio of Isolated Round Wires (5-6)

### 5.2.2.3 Reactance.

The reactance of the conductor is generally inductive and is given by the product of the radian frequency,  $\omega$ , and the self-inductance,  $L$ , of the conductor. The self-inductance of a conductor is a measure of that property which causes an opposition to a change in the current flowing in the conductor. Because skin effect redistributes the current within a conductor with changes in frequency, the inductance of the conductor does vary with frequency.

The self-inductance of a straight round wire is given (5-6) by

$$L = 0.00508\ell \left( 2.303 \log \frac{4\ell}{d} - 1 + \mu_r \kappa \right) \mu\text{H} \quad (5-9)$$

where  $\ell$  is the length in inches,  $d$  is the diameter in inches, and  $\kappa$  is a skin effect correction factor which may be determined (for copper) from Figure 5-3. For  $\ell$  and  $d$  in centimeters, Equation 5-9 becomes

$$L = 0.002\ell \left( 2.303 \log \frac{4\ell}{d} - 1 + \mu_r \kappa \right) \mu\text{H} \quad (5-10)$$

For materials other than copper,  $\kappa$  can be obtained from Figure 5-3 by using  $f' = f(\rho_c/\rho)$  instead of the actual frequency  $f$ , where  $\rho$  is the resistivity of the material and  $\rho_c$  is the resistivity of copper. For low frequencies where the current flow can be assumed to be uniform across the conductor cross-section, the inductance of a round straight wire of length  $\ell$ , diameter  $d$ , and relative permeability  $\mu_r$  (if surrounded by air) is

$$L_{LF} = 0.00508\ell \left( 2.303 \log \frac{4\ell}{d} - 1 + \frac{\mu_r}{4} \right) \mu\text{H} \quad (5-11)$$

where all the dimensions are in inches. As the frequency increases, a limiting value of inductance,  $L_{HF}$ , is approached:

$$L_{HF} = 0.00508\ell \left( 2.303 \log \frac{4\ell}{d} - 1 \right) \mu\text{H} \quad (5-12)$$

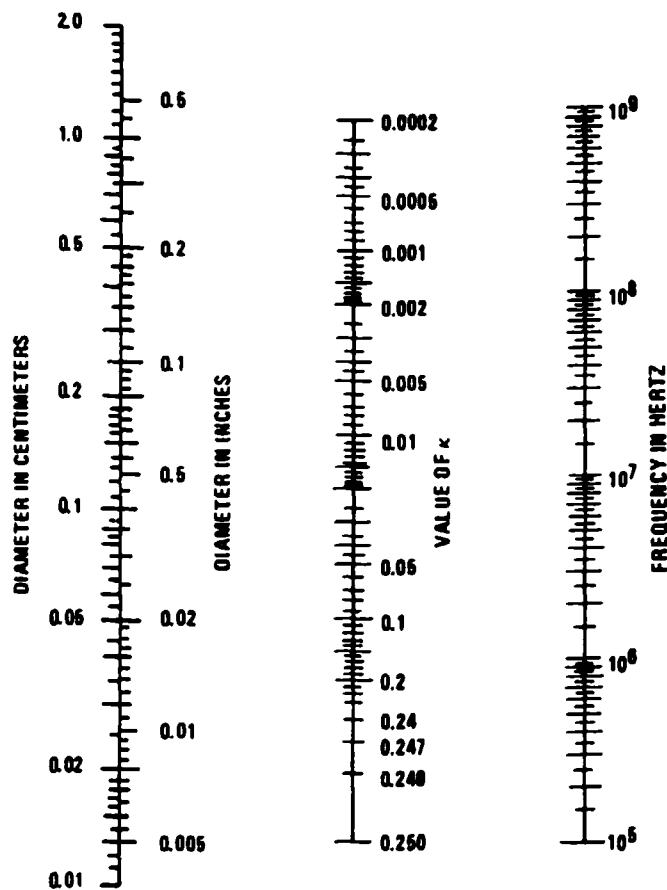


Figure 5-3. Nomograph for the Determination of Skin Effect Correction Factor (5-6)

In Equations 5-11 and 5-12, the constant 0.00508 becomes 0.002 when  $\ell$  and  $d$  are in cm.

Figure 5-4 gives the value of  $L_{LP}$  for a 1/0 AWG solid round copper conductor as a function of length, and  $L_{HP}$  for various wire lengths and diameters is given in Figure 5-5.

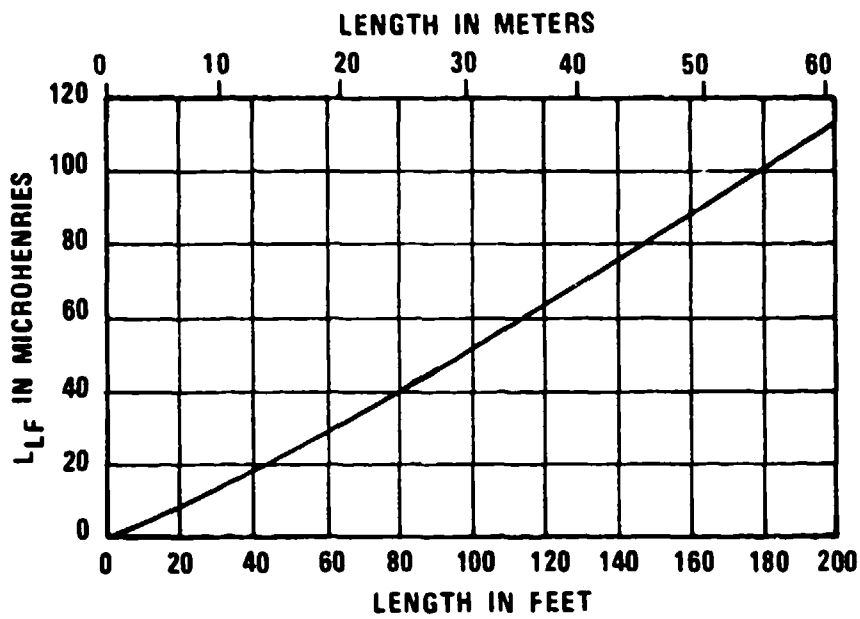


Figure 5-4. Low Frequency Self-Inductance versus Length for 1/0 AWG Straight Copper Wire (5-7)

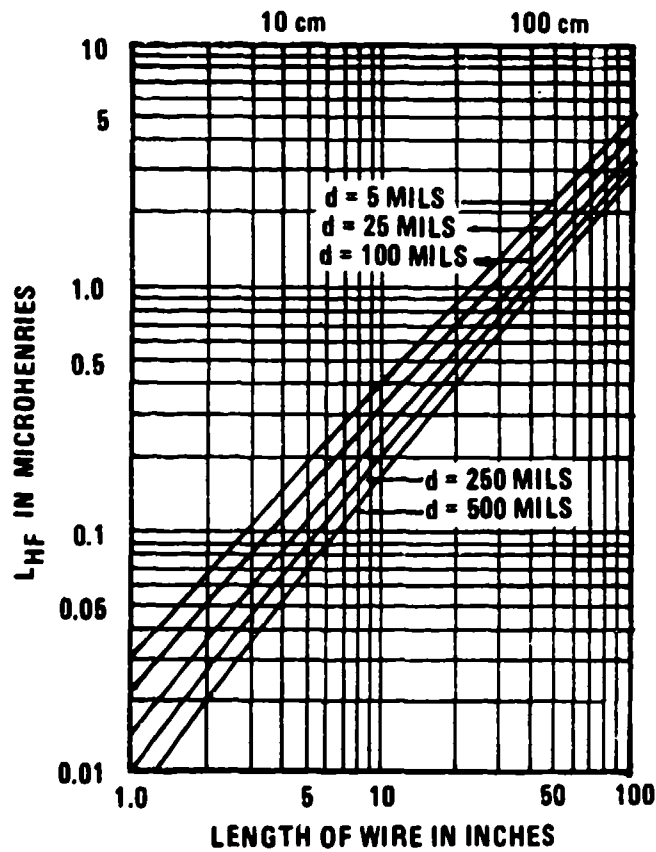


Figure 5-5. Self-Inductance of Straight Round Wire at High Frequencies (5-6)

**5.2.2.4 Proximity Effect.** When two or more conductors are in close proximity, the current flowing in one conductor is redistributed because of the magnetic field produced by the current in the other conductor. This effect is an extension of skin effect and is called proximity effect. The proximity effect tends to increase the ac resistance of a conductor to a value greater than that due to simple skin effect.

### **5.2.3 Resistance Properties vs Impedance Properties.**

Although skin effect exists at all frequencies, it becomes more significant as the frequency increases. The reactance of a conductor also increases with frequency to further increase the conductor impedance above its dc value. To design an effective ground system one must consider the relative effects of the dc resistance, the ac resistance, and the inductance upon the total impedance of a ground conductor.

Using Equation 5-1, the dc resistance of round wire conductors can be calculated. The dc resistance per 1000 feet for four standard size copper cables is given in Table 5-3. Table 5-4 gives the dc resistance and (for 60 Hz) the ac resistance, the inductance and the total impedance of various size and length conductors as determined from Table 5-3 and from Equation 5-12. At a frequency of 1 MHz, these same characteristics for 30-meter (100-foot) lengths are given in Table 5-5 as calculated from Equations 5-3 and 5-12. Note that for the larger wires (No. 2 AWG or larger) the inductance of the long (> 100 feet) cables determines the magnitude of the impedance. Also note that for the same length cables there is not as much difference in the impedance magnitudes of a small and a large cable as there is in the resistance of the two cable sizes. For example, the ratio of the dc resistance of a 30-meter (100-foot) length of No. 12 AWG copper cable to the dc resistance of a 30 meter (100 feet) of 1/0 AWG copper cable is  $0.15880/0.0098 = 16.20$ . Since the ac resistance at 60 Hz is approximately the same as the dc resistance, the ratio of the 60 Hz ac resistance of the two cables is also 16.20. At a frequency of 1 MHz the ratio of the ac resistance becomes  $1.23/0.307 = 4.01$ . However, the 60 Hz impedance ratio is only  $0.1605/0.0226 = 7.10$  and the 1 MHz impedance ratio is only  $382.85/329.49 = 1.16$ . These ratios are tabulated in Table 5-6 for comparison. From Tables 5-3 through 5-6 and the above example, the following conclusions are made:

- a. Because of the inductance, the advantages offered by a large cable such as 1/0 AWG are less than they might appear to be from a comparison of the dc resistance values.
- b. The advantage offered by a large cable, e.g., 1/0 AWG, will be somewhat more pronounced for relatively short conductor lengths than for long conductor runs. This is true because inductance increases more rapidly with length than does resistance (see Equations 5-1 and 5-9).
- c. Because of the lack of dramatic improvement in ac impedance of large cables over smaller cable sizes for long runs, consideration of materials and labor costs are relatively important and may be the deciding factor.
- d. Since even 1/0 AWG cables exhibit impedances from 22.6 mΩ to 115.8 mΩ for lengths of 30 meters (100 feet) and 137 meters (450 feet), respectively, the control of stray currents should be an essential objective in any signal grounding system.

Table 5-3

## DC Parameters of Some Standard Cables

<u>Size</u> (AWG)	<u>Diameter</u> (mils)	<u>DC Resistance</u> (Ohms/1000 ft)
No. 12	80.81	1.588
No. 8	128.5	0.6282
No. 2	257.6	0.1563
1/0	324.9	0.09827

Table 5-4

## Sixty-Hertz Characteristics of Standard Cables

<u>Size</u> (AWG)	<u>Length</u> (Ft)	<u>R<sub>ac</sub></u> ( $\Omega$ )	<u>L</u> ( $\mu$ H)	<u>X<sub>L</sub></u> ( $\Omega$ )	<u>Z</u> ( $\Omega$ )
No. 12	30	0.04764	16.532	0.00623	0.0480
No. 12	100	0.15880	62.447	0.02354	0.1605
No. 8	30	0.01885	15.684	0.00591	0.0197
No. 2	30	0.00469	14.411	0.00543	0.0072
No. 2	100	0.01563	55.379	0.02088	0.0261
No. 2	150	0.02344	86.777	0.03271	0.0402
1/0	30	0.00294	13.987	0.00527	0.0060
1/0	100	0.00980	53.964	0.0226	0.0060
1/0	150	0.01470	84.654	0.03191	0.0351
1/0	300	0.02940	181.987	0.06861	0.0746
1/0	450	0.04410	284.105	0.10710	0.1158

Table 5-5

## One-Megahertz Characteristics of Standard Cables

<u>Size</u> AWG	<u>Length</u> (Ft)	<u>R<sub>dc</sub></u> (Ω)	<u>R<sub>ac</sub></u> (Ω)	<u>L</u> (μH)	<u>X<sub>L</sub></u> (Ω)	<u>Z</u> (Ω)
No. 12	100	0.1588	1.23	60.9	382.65	382.65
No. 2	100	0.0156	0.387	53.8	338.03	338.03
1/0	100	0.0098	0.307	52.44	329.49	329.49

Table 5-6

## Impedance Comparisons Between No. 12 AWG and 1/0 AWG

<u>Frequency</u>	<u>Length</u>	$\frac{R_{dc} \text{ (No. 12 AWG)}}{R_{dc} \text{ (1/0 AWG)}}$	$\frac{R_{ac} \text{ (No. 12 AWG)}}{R_{ac} \text{ (1/0 AWG)}}$	$\frac{Z \text{ (No. 12 AWG)}}{Z \text{ (1/0 AWG)}}$
60 Hz	30 ft.	16.20	16.20	9.23
60 Hz	100 ft.	16.20	16.20	7.99
1 MHz	100 ft.	16.20	4.01	1.16

5.2.4 Effects of Geometry. Many conductor shapes can be used in the signal ground network. As is the case for the solid round conductor, the impedance of other configured conductors is dependent upon the current distribution in the conductor and hence upon the signal frequency.

#### 5.2.4.1 Stranded Cable.

A stranded cable consists of a number of wires in close proximity twisted about each other; it is more flexible than a solid conductor of the same cross-sectional area. Because of the close proximity of the wires, the skin effect within the cable redistributes most of the current to outer wires. These outer wires are in the form of a coil (due to the lay of the strand), thus increasing the self-inductance of the cable. Skin effect also increases the ac resistance as the frequency is increased.

For a given cable size, both the ac resistance and the self-inductance of a stranded conductor are greater than those of a solid round conductor. Because of their ineffectiveness at higher frequencies, it has been recommended that stranded cables not be used at frequencies over 1200 Hz (5-7). However, in many situations, large cables are required to safely carry currents produced by power faults and lightning discharges; in addition, solid wires larger than approximately 0.6 cm (0.25 in.) may be difficult to obtain.

#### 5.2.4.2 Rectangular Conductors.

At frequencies high enough to make the skin effect noticeable, the resistance ratio of a flat rectangular conductor will be lower than that of a solid round wire with the same cross-sectional area if the width-to-thickness ratio exceeds approximately 2:1. The resistance ratios for several sizes of nonmagnetic ( $\mu_r = 1$ ) rectangular conductors are plotted in Figure 5-6.

The self-inductance at lower frequencies of a rectangular conductor is (5-6)

$$L_{LF} = 0.00508 \ell \left( 2.303 \log \frac{2\ell}{b+c} + 0.5 + .2235 \frac{b+c}{\ell} \right) \mu H, \quad (5-13)$$

where  $\ell$  is the length,  $b$  is the width and  $c$  is the thickness, and all the dimensions are in inches. For the dimensions in cm, Equation 5-13 is

$$L_{LF} = 0.002 \ell \left( 2.303 \log \frac{2\ell}{b+c} + 0.5 + 0.2235 \frac{b+c}{\ell} \right) \mu H. \quad (5-14)$$

If  $\ell$  is larger than 50 ( $b+c$ ), the last term in each equation may be neglected.

The sharp edges on rectangular conductors tend to radiate energy into space and a flat conductor may become an efficient antenna. To reduce the efficiency of the antenna and minimize this radiation, the edges of the rectangular conductor can be rounded to form an elliptical shape.

#### 5.2.4.3 Tubular Conductors.

Tubular conductors provide the best compromise between factors such as availability, cost weight, cross-sectional area, skin effect, resistance ratio and inductance. By using the actual cross-sectional area of the conductive material, the dc resistance of tubular conductors can be determined from Equation 5-1; it is given for three different diameter copper tubes in Figure 5-7.



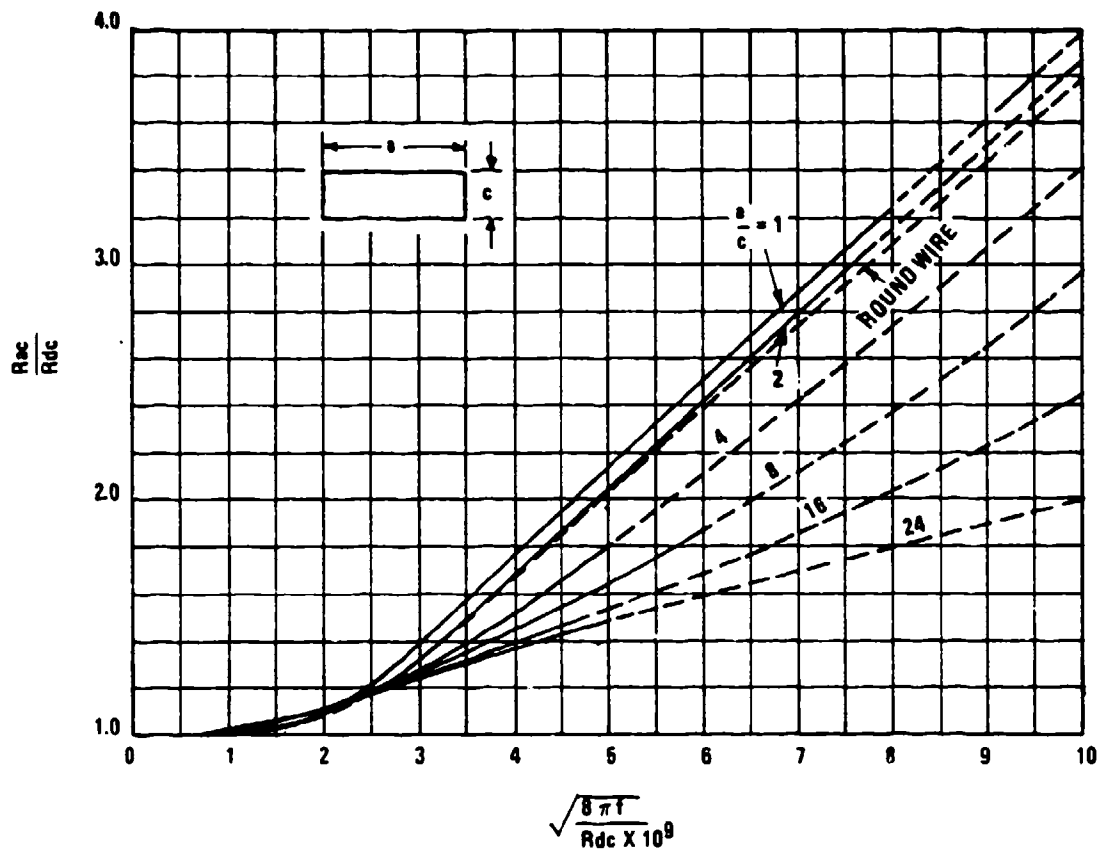


Figure 5-6. Resistance Ratio of Rectangular Conductors (5-3)

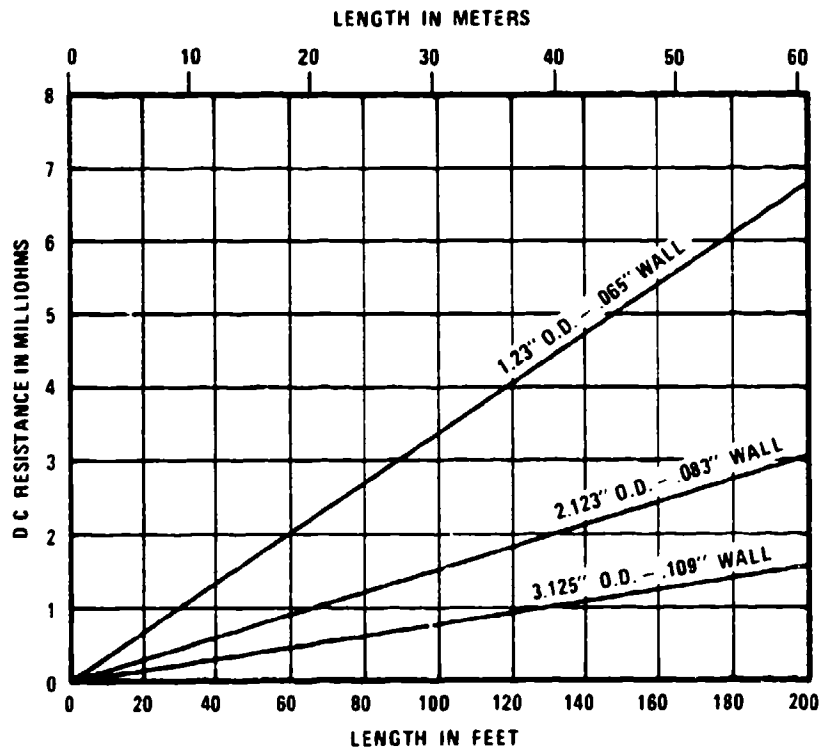


Figure 5-7. Resistance versus Length for Various Sizes of Copper Tubing (5-7)

Both the dc and the ac resistances of a tubular conductor are greater than those of a solid conductor with the same outside diameter. However, the ac resistance does not increase as much as the dc resistance, and, therefore, the resistance ratio of a tubular conductor is always less than that of a solid conductor. The ac resistance for four sizes of copper tubing is given in Figure 5-8, and the resistance ratio for isolated nonmagnetic tubular conductors of various sizes is given in Figure 5-9. For a given length of conductor, the ac resistance per unit weight (i.e., per given amount of copper) is less at high frequencies for tubular conductors than for any other shape.

The self-inductance of a conductor is reduced by the absence of a conductive medium in the center (5-7). Therefore, the self-inductance of a tubular conductor will be less than that of a solid conductor with the same diameter. The self-inductance of three representative sizes of copper tubes is given in Figure 5-10.

**5.2.4.4 Structural Steel Members.** The steel I-beam in the structural framework of a building is another conductor that is frequently used as a ground conductor. The resistivity of steel is approximately ten times that of copper; however, the skin depth of steel is greater than 3 times that of copper. This increased skin depth in steel increases the conducting area for high frequency currents. For example, in comparing a 0.3 meter (12-inch) I-beam with a 4/0 AWG copper cable, the perimeter of the I-beam is about 30 times as great and with a factor of 3 increase in the skin depth, the conducting area for high frequency currents in the steel I-beam is close to 90 times larger. This advantage is offset somewhat by the fact that the current tends to flow in the edges of the I-beam and by the surface roughness. The ac resistance will be increased by a factor of 4 because of this surface roughness and current distribution. Even so, the ac resistance of a 4/0 AWG copper cable is 4.25 times as great as that of a 0.3 meter (12-inch) I-beam. In addition, the building framework usually offers many paths in parallel, thus lowering both the ac resistance and the inductance between any two points (5-8).

**5.3 SIGNAL REFERENCE SUBSYSTEM NETWORK CONFIGURATIONS.** Within a piece of equipment the signal reference subsystem may be a sheet of metal which serves as a signal reference plane for some or all of the circuits in that equipment. Between equipments, where units are distributed throughout the facility, the signal ground network usually consists of a number of interconnected wires, bars or a grid which serves an equipotential plane. Whether serving a collection of circuits within an equipment or serving several equipments within a facility, the signal reference subsystem will be a floating ground, a single-point ground, or a multiple-point ground known as a multipoint or equipotential plane. Of the aforementioned signal reference subsystems, the equipotential plane is the optimum ground for communications-electronics facilities. For existing facilities where the presence of equipment prohibit the installation of an equipotential plane beneath, on, or in the floor, the plane may be installed overhead and the equipment connected to it. It is desirable, but not mandatory, to retrofit existing C-E facilities with equipotential planes.

#### **5.3.1 Floating Ground.**

A floating ground is illustrated in Figure 5-11. In a facility, this type of signal ground system is electrically isolated from the building ground and other conductive objects. Hence, noise currents present in the building's ground system will not be conductively coupled to the signal circuits. The floating ground system concept is also employed in equipment design to isolate the signal returns from the equipment cabinets and thus prevent noise currents in the cabinets from coupling directly to the signal circuits.

The effectiveness of floating ground systems depends on their true isolation from other nearby conductors, i.e., to be effective, floating ground systems must really float. In large facilities, it is often difficult to achieve a completely floating system, and even if complete isolation is achieved it is difficult to maintain such a system (5-9).

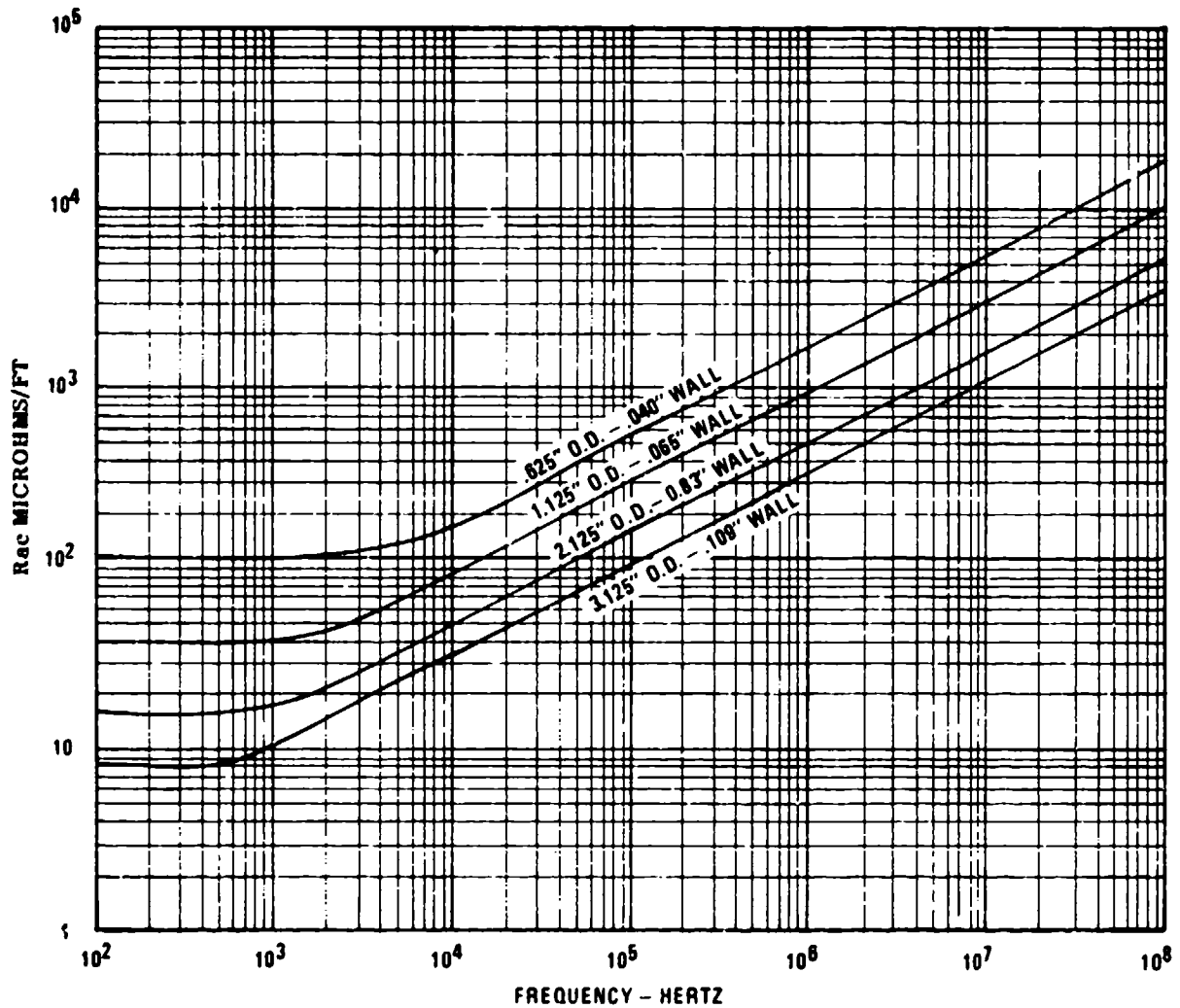


Figure 5-8. AC Resistance versus Frequency for Copper Tubing (5-7)

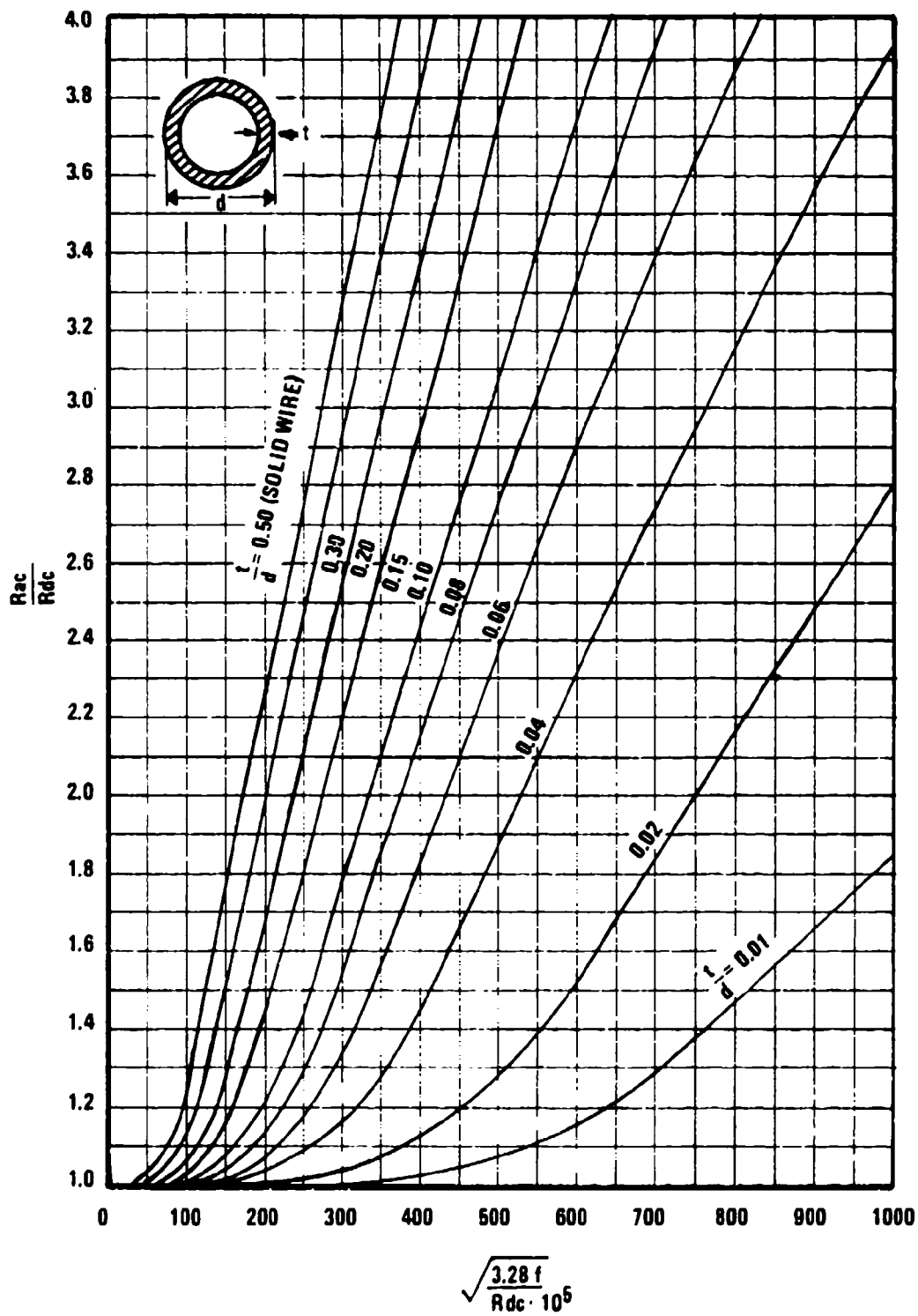


Figure 5-9. Resistance Ratio of Nonmagnetic Tubular Conductors (5-3)

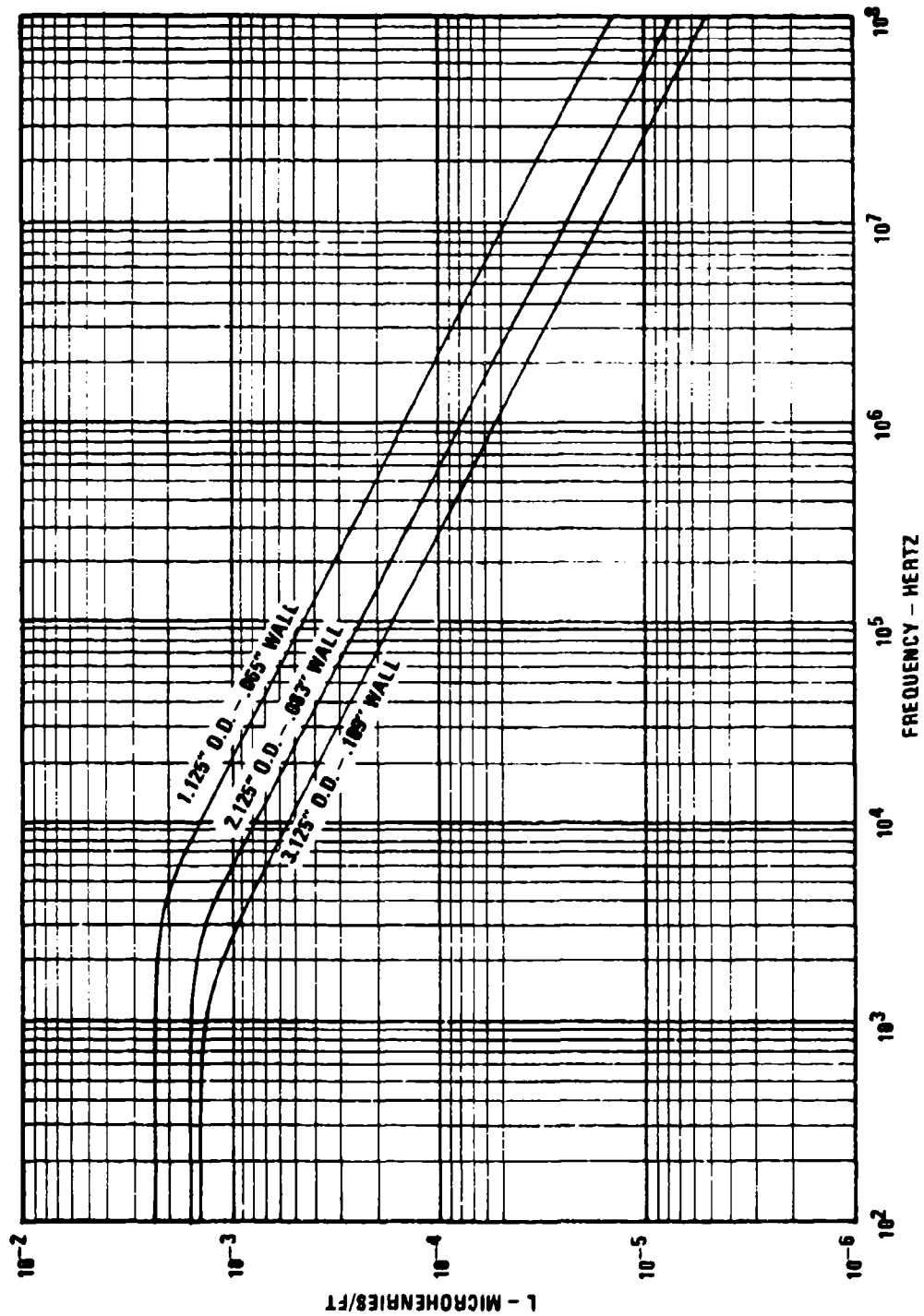


Figure 5-10. Inductance versus Frequency for Various Sizes of Copper Tubing (5-7)

In addition, a floating ground system suffers from other limitations. For example, static charge buildup on the isolated signal circuits is likely and may present a shock and a spark hazard. In particular, if the floated system is located near high voltage power lines, static buildup is very likely. Further, in most modern electronic facilities, all external sources of energy such as commercial power sources are referenced to earth grounds. Thus, a danger with the floating system is that power faults to the signal system would cause the entire system to rise to hazardous voltage levels relative to other conductive objects in the facility. Another danger is the threat of flashover between the structure or cabinet and the signal system in the event of a lightning stroke to the facility. Not being conductively coupled together, the structure could be elevated to a voltage high enough relative to the signal ground to cause insulation breakdown and arcing. This system generally is not recommended for C-E facilities.

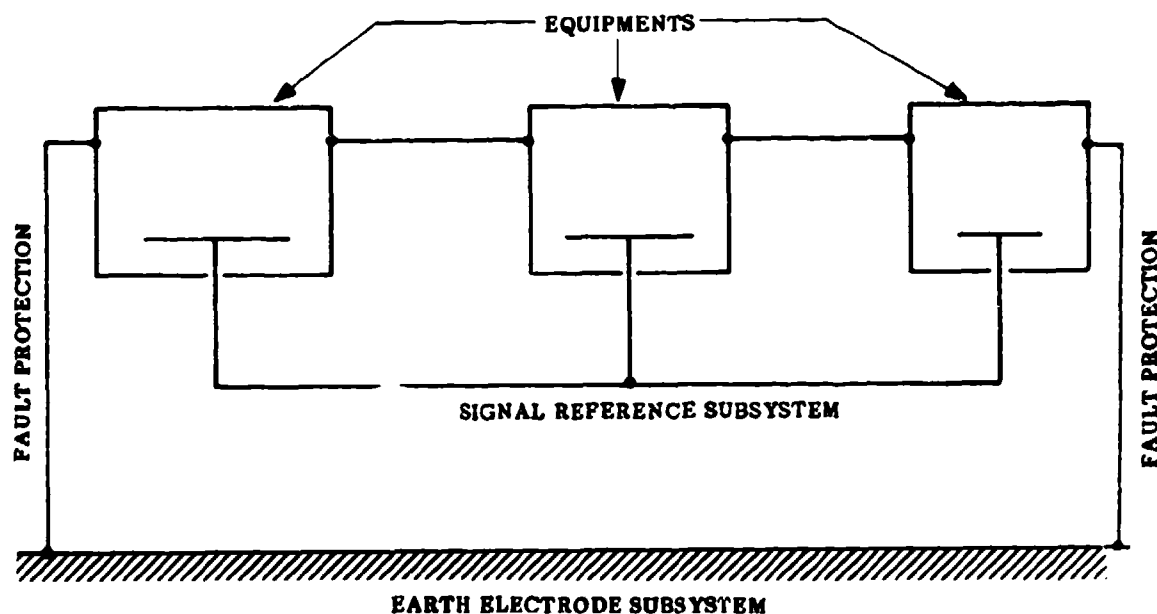


Figure 5-11. Floating Signal Ground

### 5.3.2 Single-Point Ground. (For lower frequencies, 0-30 kHz up to 300 kHz)\*

A second configuration for the signal ground network is the single-point approach illustrated in Figure 5-12. With this configuration, the signal circuits are referenced to a single point, and this single point is then connected to the facility ground. The ideal single-point signal ground network is one in which separate ground conductors extend from one point on the facility ground to the return side of each of the numerous circuits

\* Refer to 5.4.3 for definition of frequency limits.

located throughout a facility. This type of ground network requires an extremely large number of conductors and is not generally economically feasible. In lieu of the ideal, various degrees of approximation to single-point grounding are employed.

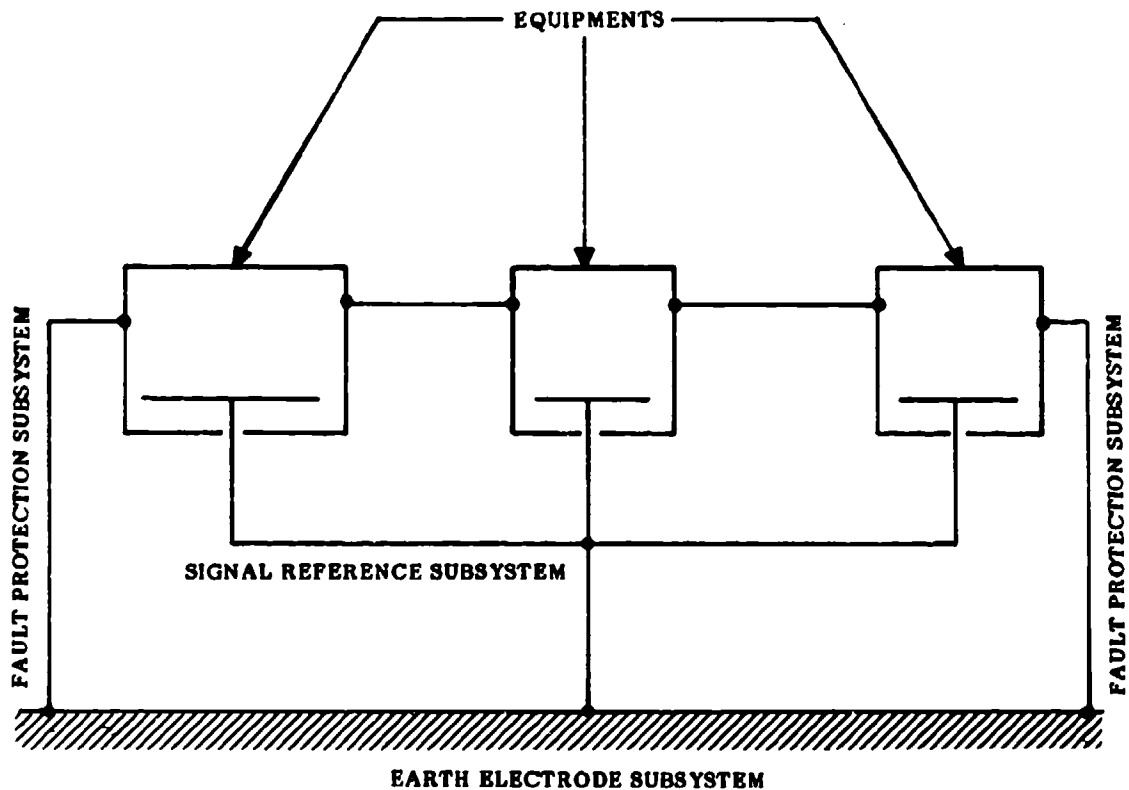


Figure 5-12. Single-Point Signal Ground (For Lower Frequencies)

The configuration illustrated by Figure 5-13 closely approximates an ideal single-point ground. It uses individual ground buses extending from an earth electrode subsystem to each separate electronic system. In each system, the various electronic subsystems are individually connected at only one point to this ground bus. Another frequently used approximation to the ideal is illustrated in Figure 5-14. Here the ground bus network assumes the form of a tree. Within each system, each subsystem is single-point grounded. Each of the system ground points is then connected to a tree ground bus with a single insulated conductor (usually yellow).

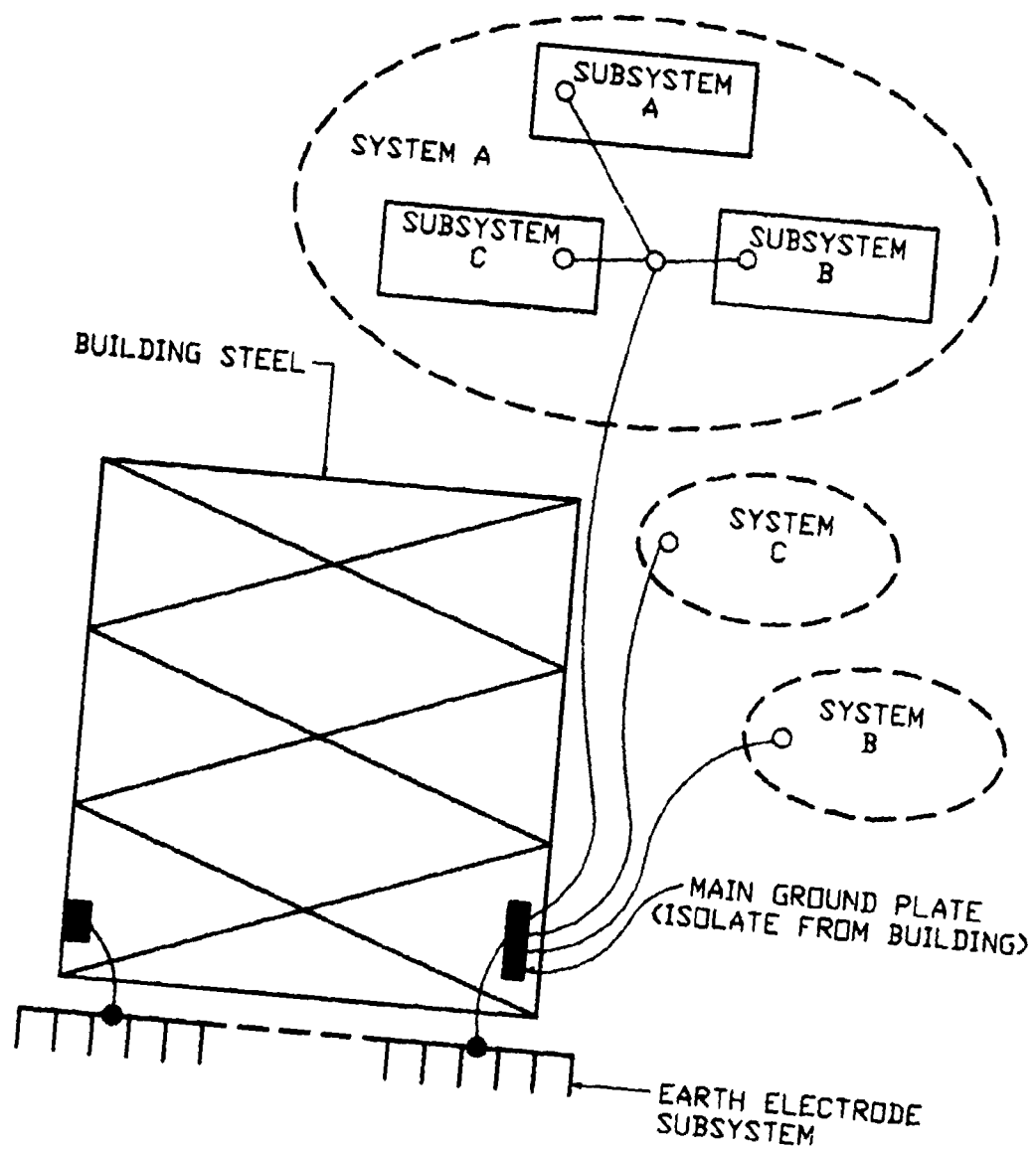


Figure 5-13. Single-Point Ground Bus System Using Separate Risers (Lower Frequency)



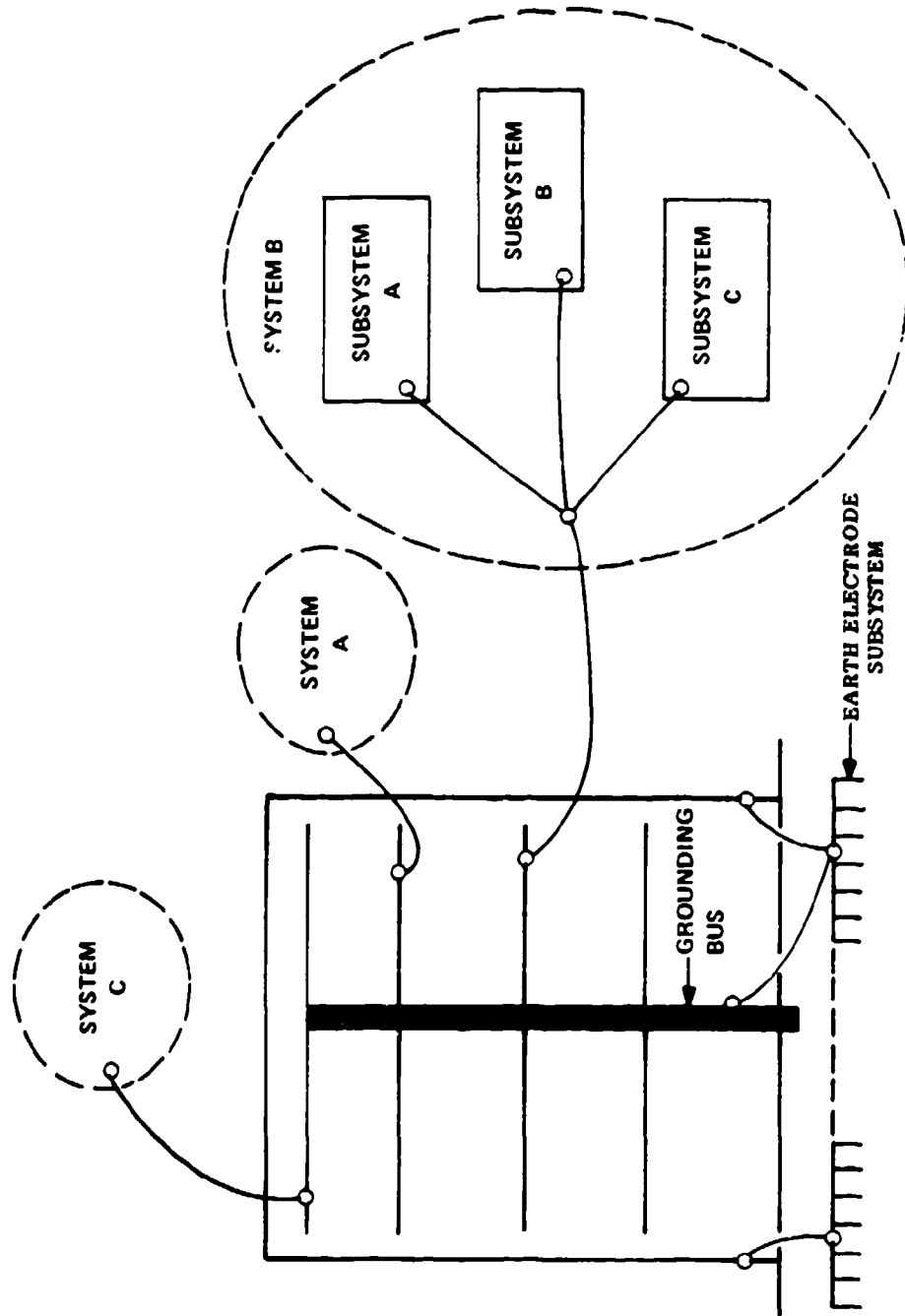


Figure 5-14. Single-Point Ground Bus System Using a Common Bus

The single-point ground accomplishes each of the three functions of signal circuit grounding mentioned at the beginning of this chapter. That is, a signal reference plane is established in each unit or piece of equipment and these individual reference planes are connected together and to the earth electrode subsystem. An important advantage of the single-point configuration is that it helps control conductively-coupled interference. As illustrated in Figure 5-15, closed paths for noise currents in the signal ground network are avoided, and the interference voltage,  $V_N$ , in the facility ground system is not conductively coupled into the signal circuits via the signal ground network. Therefore, the single-point signal ground network minimizes the effects of lower frequency noise currents which may be flowing in the facility ground.

Single-point grounds, however, also become transmission lines at higher frequencies with earth being the other side of the line. In addition, every piece of equipment bonded to this transmission line will act as a tuned stub. In the presence of digital signals (square waves) the tuned circuits will ring at the specific frequencies to which they are resonant. Since single-point grounds behave as transmission lines at rf frequencies, they will have different impedances as a function of frequency, i.e., they may appear as inductors, capacitors, tuned circuits, insulators or pure resistance, and therefore become extremely poor grounds. In a large installation, another major disadvantage of the single-point ground configuration is the requirement for long conductors. The long conductors ( $1/8 \lambda$  at the highest frequency of concern) prevent the realization of a satisfactory reference for

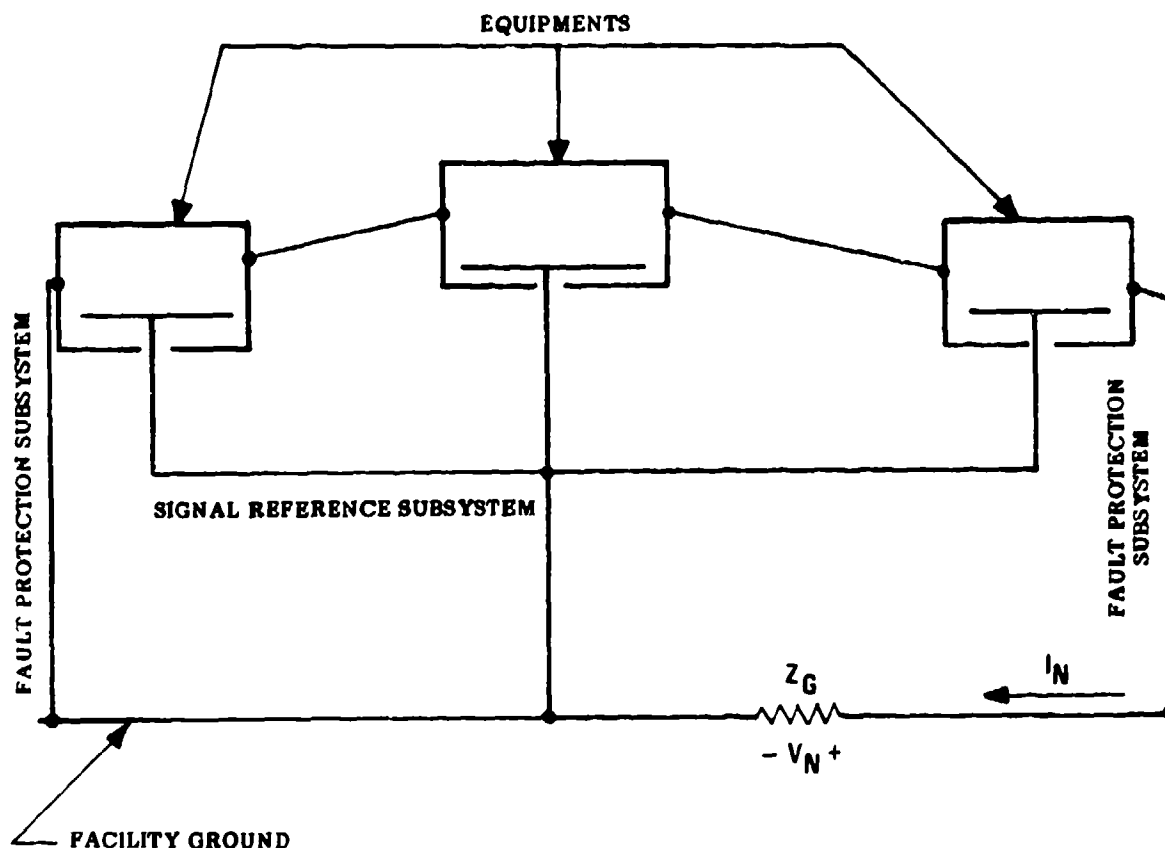


Figure 5-15. Use of Single-Point Ground Configuration to Minimize Effect of Facility Ground Currents

higher frequencies because of large self-impedances. Further, because of stray capacitance between conductors, single-point grounding essentially ceases to exist as the signal frequency is increased (5-10). Because of the aforementioned reasons, single-point grounds are not recommended for use in communications electronics facilities.

### 5.3.3 Multipoint Ground. (For higher frequencies, 30-300 kHz and above)

The multipoint ground illustrated in Figure 5-16 is the third configuration used for signal ground networks. The multipoint ground utilizes many conductive paths from the earth electrode subsystem to various electronic systems or subsystems within the facility. Within each subsystem, circuits and networks are multiply connected to this ground network. Thus, in a facility, numerous parallel paths exist between any two points in the ground network as shown in Figure 5-17.

Multipoint grounding frequently simplifies circuit construction inside complex equipments; it is the only realistic method for the grounding of higher frequency signal circuits. This method of grounding permits equipments employing coaxial cables to be more easily interfaced since the outer conductor of the coaxial cable does not have to be floated relative to the equipment cabinet or enclosure. The multipoint grounding has the disadvantage of exhibiting transmission line characteristics at rf frequencies. To be effective, a multipoint ground system requires an equipotential ground plane whenever the conductors exceed  $1/8 \lambda$  at the highest frequency of concern (5-11).

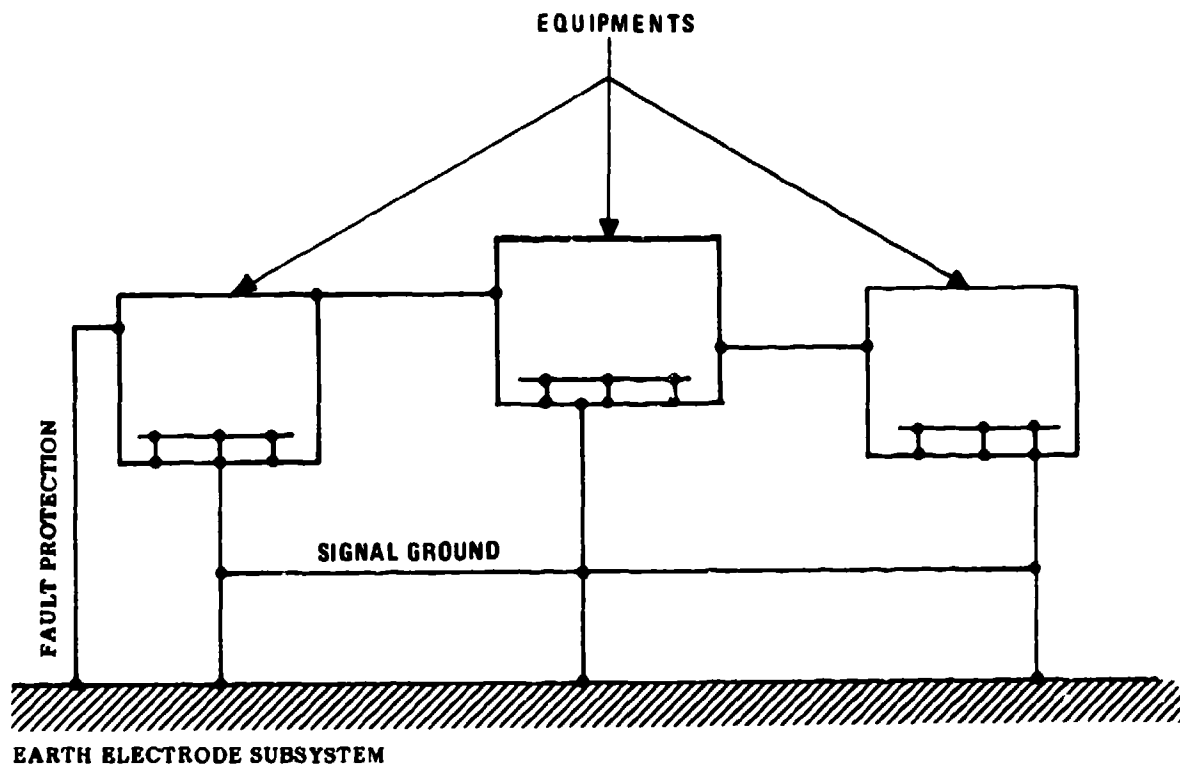


Figure 5-16. Multipoint Ground Configuration

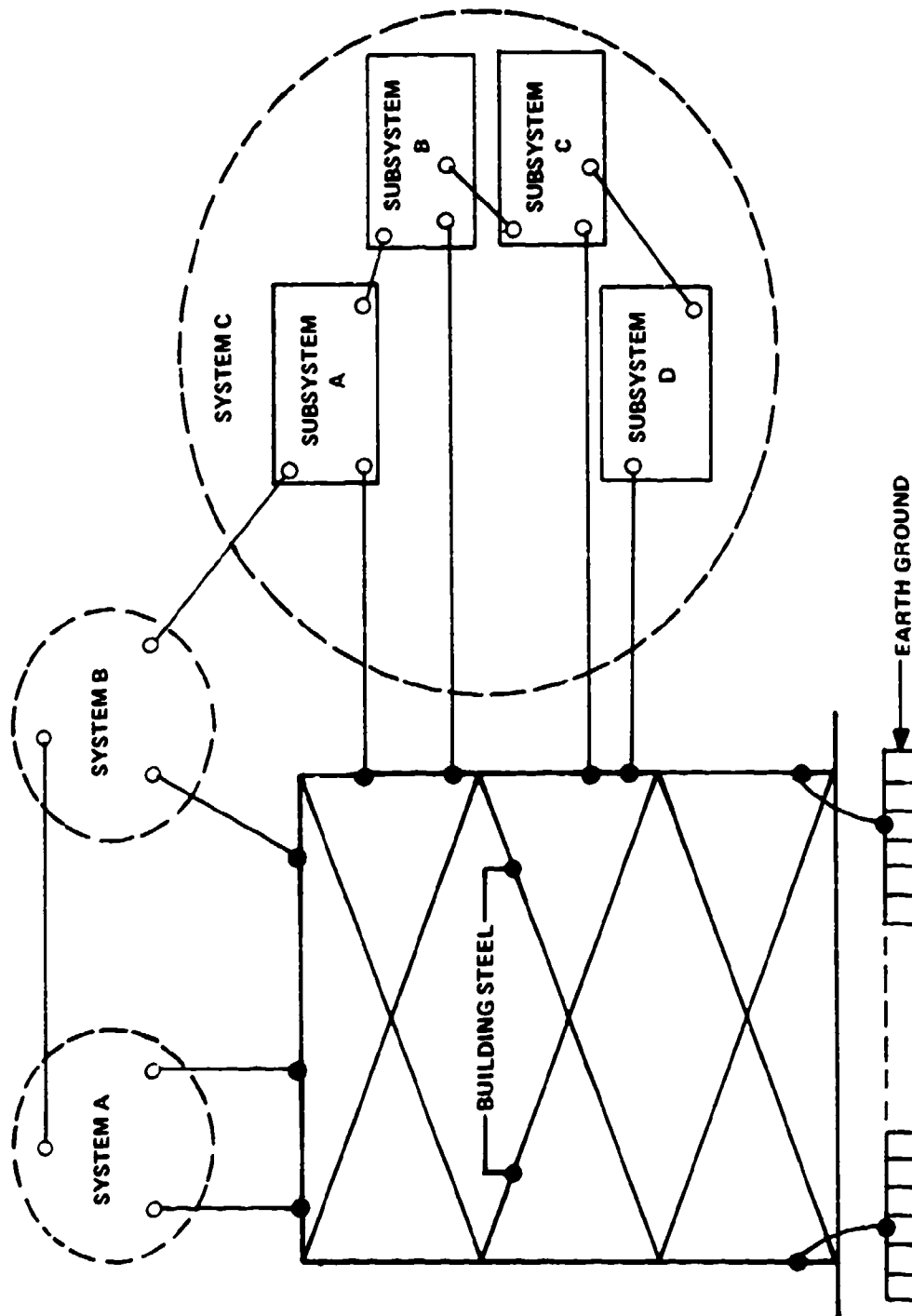


Figure 5-17. Use of Structural Steel in Multiple-Point Grounding

Care must also be taken to ensure sixty hertz power currents and other high amplitude lower frequency currents flowing through the facility ground system do not conductively couple into signal circuits and create intolerable interference in susceptible lower frequency circuits.

#### 5.3.3.1 Equipotential Plane.

The importance of equipotential ground planes cannot be overemphasized for proper equipment operation, as well as for EMI and noise/static suppression. An equipotential ground plane implies a mass, or masses of conducting material which, when bonded together, offers a negligible impedance to current flow. Connections between conducting materials which offer a significant impedance to current flow, can place an equipotential plane at a high potential with respect to earth. High impedance interconnections between metallic members subject to large amounts of current due to power system faults can be extremely hazardous to personnel and equipment. The RFI effect of an equipotential plane or system must however be carefully considered, and it is important to understand that grounding may not, in and of itself, reduce all types of RFI. On the contrary, grounding a system may in some instances increase interference by providing conductive coupling paths or radiative or inductive loops.

Many of the deficiencies of the wire distribution system can be overcome by embedding a large conducting medium, in the floor under the equipments to be grounded. For existing facilities this system may be installed above the equipment to be grounded. A large conducting surface presents a much lower characteristic impedance than that of wire because the characteristic impedance ( $Z_0$ ) is a function of  $L/C$ , hence as capacity to earth increases,  $Z_0$  decreases. The capacity of a metallic sheet or grid to earth is much higher than that of wire. If the size of the sheet is increased and allowed to encompass more area, the capacitance increases. Also, the unit length inductance decreases with width, which further decreases  $Z_0$ . If the dimensions of a metallic sheet increase extensively (as in the case of conducting floor), the characteristic impedance approaches a very low value. In this case, the characteristic impedance would be quite low throughout a large portion of the spectrum. This, in turn, would establish an equipotential reference plane for all equipments bonded to it.

Although it is not necessary from a functional point of view, terminating the surface to an earth connection presents the following advantages:

- a. Personnel safety is not dependent on long cable runs for protection against power faults.
- b. Low impedance is provided for power and radio frequencies.

Grounding buses in a communication facility where higher frequencies are present, act as lossy transmission lines and therefore must be treated as such. Due to this phenomena single-point grounds and multipoint grounds employing ground buses are high impedance grounds at higher frequencies. To be effective at the higher frequencies, the multipoint ground system requires the existence of an equipotential ground plane. Equipotential Planes are sometimes considered to exist in a building with a metal floor or ceiling grid electrically bonded together, or in a building with the ground grid embedded in a concrete floor connected to the structural steel and the facility ground system. Equipment cabinets are then connected to the equipotential plane. Chassis are connected to the equipment cabinets and all components, signal return leads,

etc., are connected to the chassis. The equipotential plane is then terminated to the earth electrode subsystem and to the main structural steel via multiple connections, to assure personnel safety and a low impedance path for all frequencies and signals. It is again emphasized, however, that care must be taken not to create loops which can couple signals from one system to another.

The equipotential plane also offers the following additional advantages;

a. Any "noisy" cable or conductor connected to the receptor, i.e., receivers, modems, etc., through or along such a ground plane will have its field contained between the conductor and the ground plane. The noise field can be "shorted out" by filters and bond straps because the distance between these "transmission line" conductors is very small. Shorting out the noise field has the desirable effect of keeping noise current from flowing over the receptor case and along any antenna input cables.

b. Filters at the interface terminals of equipment can operate more effectively when both terminals of their equivalent "transmission line" are available. As in a, above, a large conducting surface makes it possible to contain the field carried by the offending conductor, in such a way that it can be more easily prevented from traveling further.

c. A large conducting surface may also shield or isolate rooftop antennas from noisy cables below it.

5.3.3.2 Types of Equipotential Planes. Conducting materials that can be utilized for equipotential planes are (a) a copper grid embedded in the concrete floor such as a computer floor, (b) a subfloor of aluminum, copper, phosphor bronze screen or sheet metal laid underneath the floor tile or carpet or (c) a ceiling grid above the equipment. Additional data and information on each of these planes can be found in para 1.5.1.1.1 of Vol II.

#### 5.3.4 Floating System.

The floating ground system is completely insulated from the building or from any wiring that may be a source of circulating currents. The effectiveness of floating ground systems depends on their true isolation. In large systems, it is difficult to provide required isolation to maintain a good quality floating ground. Insulation breakdown occurs easily because static charges, fault potentials and lightning potentials may accumulate between the floating ground and other accessible grounds, such as external power line neutrals, water pipes, etc. Due to the personnel hazards from the difference of potential between the floating ground and building ground, this system is not recommended.

The preferred grounding method is to have an equipotential plane bonded to the earth electrode subsystem and building structure steel at multiple points with the structural steel also bonded to the earth electrode subsystem. In those facilities which do not have structural steel, multiple copper downloads should be connected from the equipotential plane to the earth electrode subsystem.

#### 5.4 SITE APPLICATIONS.

Because of the interference threat that stray power currents present to audio, digital, and control circuits (or others whose operating band extends down to 60 Hz or below), steps must be taken to isolate these large currents from signal return paths. Obviously, one way of lessening the effects of large power currents is to configure the signal ground system so that the signal return path does not share a path common with a power return. This can be accomplished by making sure that the grounding conductor (green wire) of the power system is always run in the same cable, conduit, duct, or raceway with the phase and neutral conductors to the first service disconnect and then bonded to the earth electrode subsystem.

The first step in the development of an interference-free signal reference subsystem for an equipment or a facility is to assure that the ac primary power return lines are interconnected with the safety grounding network at only one point. Isolation of ac power returns from the signal reference subsystem is a major factor toward reducing many noise problems. Additional steps should also be taken to minimize other stray ac currents such as those resulting from power line filters. (One way of reducing these currents is to limit the number of filter capacitors in an installation by using common filtered ac lines wherever possible or by locating the filters as near as possible to the power service entry of the facility.)

To meet the safety requirements while minimizing the effects of power currents flowing with signal currents through a common impedance, a single connection\* between the power distribution neutral and the earth electrode subsystem is necessary. This single connection eliminates conductive loops in which circulating (power) currents can flow to produce interference between elements of the signal reference network.

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\*This connection to the earth electrode subsystem should be made from the first service disconnect. Care should be taken to ensure that the signal reference, fault protection, and lightning protection subsystems are bonded to the earth electrode subsystem at separate ground rod locations.

**5.4.1 Lower Frequency Network (0-30 kHz, and in some cases up to 300 kHz).** The lower frequency grounding network for the facility should conform to the following principles:

a. It should be isolated from other ground networks including structural, safety, lightning and power grounds, etc. The purpose of this isolation is to prevent stray currents (primarily 50/60 Hz power) from developing voltage differentials between points on the ground network.

b. The inter-equipment or facility ground system should not be expected to provide the primary return path for signal currents from the load to the source. Figure 5-18 illustrates a way of discriminating against those extraneous signals which may inductively or capacitively induce currents into the grounding network and develop differential voltages between the source and the load. For example, Figure 5-19 illustrates a practice that is not recommended. If only one source and one load constitute the entire system or if the various source-load pairs within the system are essentially noninterfering in nature of their operation, this grounding arrangement may be acceptable.

c. The lower frequency grounding network must be connected to the earth electrode subsystem at only one point.

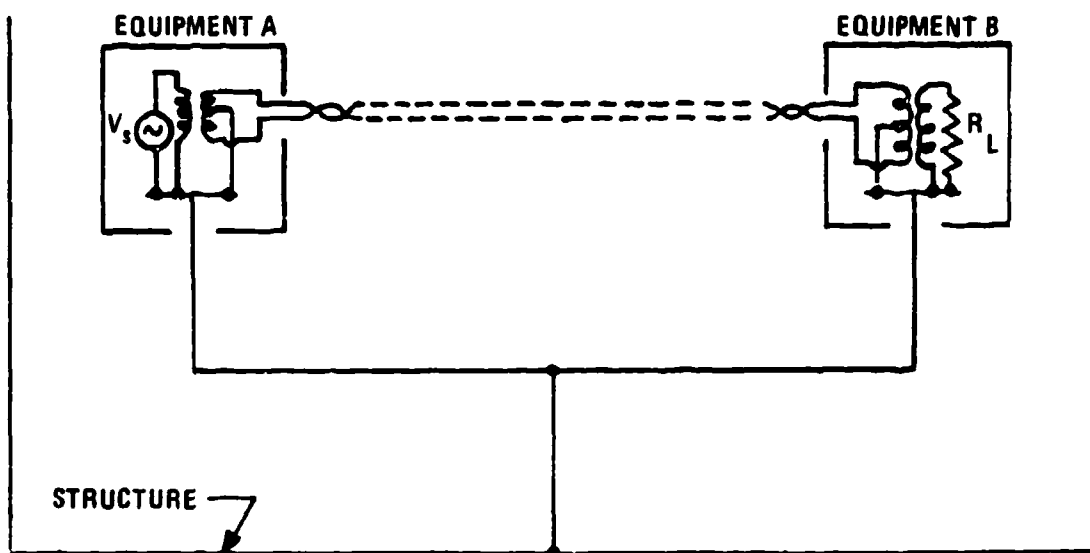


Figure 5-18. Recommended Signal Coupling Practice for Lower Frequency Equipment



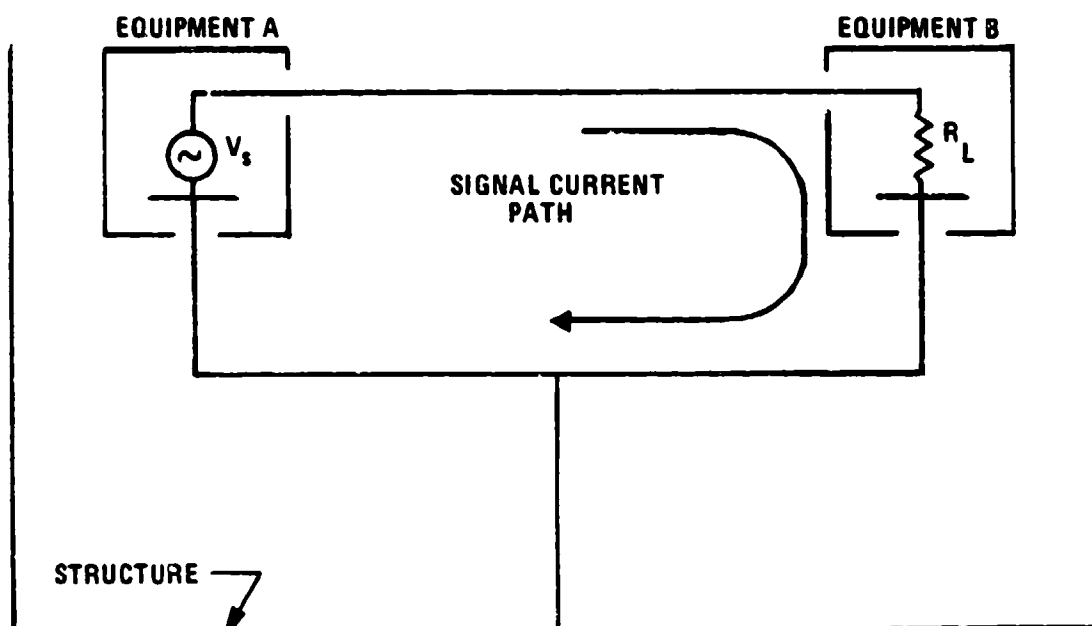


Figure 5-19. Ground Network Used as Signal Return  
(Practice Not Generally Recommended)

d. The network must be configured to minimize conductor path lengths. In facilities where the equipments to be connected to the ground network are widely separated, more than one network should be installed.

e. Finally, the conductors of the network are to be routed in a manner that avoids long runs parallel to primary power conductors, lightning down conductors, or any other conductor likely to be carrying high amplitude currents.

#### 5.4.2 Higher Frequency Network (> 300 kHz, and in some cases down to 30 kHz).

The higher frequency (equipotential) network provides an equal potential plane with the minimum impedance between the associated electronic components, racks, frames, etc. This plane shall be used at facilities or areas within facilities where interface frequencies are over 300 kHz and may be used at sites where interface frequencies are as low as 30 kHz. In higher frequency systems, equipment chassis are frequently used as the signal reference. The chassis in turn is usually connected to the equipment case at a large number of points to achieve a low impedance path at the frequencies of interest. See Para. 5.4.3.

The National Electrical Code requires that equipment cases and housings be grounded to protect personnel from hazardous voltages in the event of an electrical fault. Stray currents in the fault protection network can present an interference threat to any signal system whose operating range extends down into the lower frequency range and should be eliminated. Where such problems exist, it is advisable to attempt to reduce the impedance of the reference plane as much as possible. A practical approach is to interconnect equipment enclosures with the equipotential plane, via building structural steel, cable trays, conduit, heating ducts, piping, etc., into the earth electrode subsystem to form as many parallel paths as possible. It should be recognized that because of the inductance and capacitance of the network conductors, such multipoint ground systems offer a low impedance only to the lower frequency noise currents; however, these currents can be the most troublesome in many facilities. Higher frequencies find a much lower impedance to ground through the distributed capacity of the equipotential plane.

#### 5.4.3 Frequency Limits.

The question remaining concerns the frequency below which signals can be considered as lower frequency. Certainly the dividing line between the lower and higher frequency should be high enough to include all audio communications signals. Since digital systems employ frequencies which extend from dc up to several hundred MHz, a decision based on pulsed-signal considerations is more appropriate. To minimize the possibility that the ground bus conductors will form antennas, the lengths should not exceed 0.02 wavelength which is approximately 21 meters (70 feet) at 300 kHz. Since the grounding buses in medium to large sized facilities may extend 21 meters (70 feet), 300 kHz appears to be the maximum frequency for which a single-point grounding system should be used. At frequencies up to 30 kHz, conductor lengths up to 210 meters (700 feet) can be approached without exceeding the 0.02 wavelength criteria.

MIL-STD-188-124A establishes the lower frequency network range from dc to 30 kHz and in some cases (depending on the interface frequency) up to 300 kHz. The higher frequency network range extends above 300 kHz and may in some cases be used at sites where the interface frequencies are as low as 30 kHz. The frequency range from 30 kHz to 300 kHz is a mutual area and may be considered as either higher or lower depending upon the interface frequency.

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## CHAPTER 6

## INTERFERENCE COUPLING AND REDUCTION

6.1 INTRODUCTION.

A large number of diverse equipments are usually present in an electronics complex. The various systems and subsystems making up the complex may be concentrated in a small area such as a single room or they may be distributed over a wide geographical area and be located in several buildings. Whether the distances between individual equipments are large or small, the entire system must function as an integral unit. Each equipment must supply its designated output -- whether it be audio or rf, or analog or digital -- to some terminal point such as an antenna, land line, or another piece of equipment. Both primary and backup power must be supplied. Critical points in the system must be monitored both locally and remotely. To perform all the required tasks and functions, many control, power distribution, and signal transmission networks are necessary.

Within the interconnected complex, many potentially incompatible signals are present. For example, at one extreme are the large power sources (primarily dc and 60 Hz) supplying the various subsystems. At the other extreme, low level dc and very low frequency signals from monitors, indicators and other specialized devices are present. Also in the low frequency range are audio signals used for communications and control functions. In the higher frequency region of the spectrum are the rf signals ranging from hf to microwaves used for communications, surveillance and tracking, and other functions. These signals range in amplitude from the microwatt levels typical at communications receiver inputs to the kilowatt and megawatt levels transmitted by some radar systems. Ranging from audio frequencies into the rf region are the broadband display and communications systems, both analog and digital, which may span from a few hertz to several megahertz in frequency and may range in amplitude from a few millivolts to a few volts. Falling in overlapping frequency ranges, these various signals present within the complex may interact in an undesirable manner to cause interference (generally manifested as annoying "noise").

Interference is any extraneous electrical or electromagnetic disturbance that tends to interfere with the reception of desired signals or that produces undesirable responses in electronic systems. Interference can be produced by both natural and man-made sources either external or internal to the electronic system. The major objective of interference reduction in modern electronic equipments and facilities is to minimize and, if possible, prevent degradation in the performance of the various electronic systems by the interactions of undesired signals, both internal and external.

The correct operation of complex electronic equipments and facilities is inherently dependent on the frequencies and amplitudes of both the signals utilized in the system and the interference signals present in the facility. If the frequency of an undesired signal is within the operating frequency range of the system, errors in the system response may be obtained. The extent of the system response is a function of the amplitude of the undesired signal relative to that of the desired signal. For example, in systems operating with high level signals, undesired signals with amplitudes on the order of volts may be tolerable, while in low level systems a few microvolts may produce intolerable errors in the response of the system. An important element in the control of unwanted interactions between signals is the proper grounding of the system.

An ideal signal system is a simple signal generator-load pair as shown in Figure 6-1. With no extraneous voltages present within the loop, this simple pair is free of interference. Consider, however, what happens when the current return path is non-ideal and sources of noise are present as shown in Figure 6-2. Unless noise voltages  $V_{N1}$  and  $V_{N2}$  are identical, a voltage difference will exist between the low side of the generator (Node 1) and the low side of the load (Node 2). As shown in Figure 6-3, this voltage difference effectively appears in the signal transfer loop in series with the signal generator and produces noise currents in the load. Four ways of combating this noise problem are as follows:

- a. Isolate the source-load pair from the noise sources; i.e., float the system and provide the necessary shielding and filtering to prevent coupling by other means.
- b. Connect the low side of the loop to the reference plane at either Node 1 or Node 2 but not at both.
- c. Reduce the impedance,  $Z_{\text{return}}$ , of the path connecting the two noise sources.
- d. Reduce the magnitudes of  $V_{N1}$  and  $V_{N2}$  through the control of the currents producing them by lowering the impedance through which these currents flow.

Practical electronics circuits typically are a collection of several source-load combinations such as shown in Figure 6-4. These various source-load combinations may be functionally dependent on each other. Hence each individual source-load pair can not operate in isolation; there must be coupling between pairs. For example, one source may be driving several loads; one load may be receiving signals from several sources; or the load for one signal source may serve as the source for another load. At the circuit level, numerous sources and loads are connected in an interrelated fashion and the use of individual return paths for each source-load pair becomes impractical. It is more realistic to establish a common ground or reference plane which serves as the return path for several signals. The control of undesired network responses, particularly in high gain and/or higher frequency circuits, often requires the establishment of a common signal reference to which functional grouping of components, circuits, and networks can be connected. Ideally, this common reference connection offers zero impedance paths to all signals for which it serves as a reference. The several signal currents within the network can then return to their sources without creating unwanted conductive coupling between circuits.

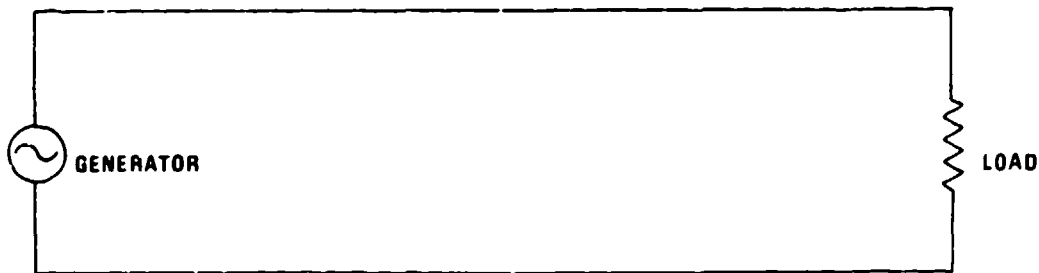


Figure 6-1. Idealized Energy Transfer Loop

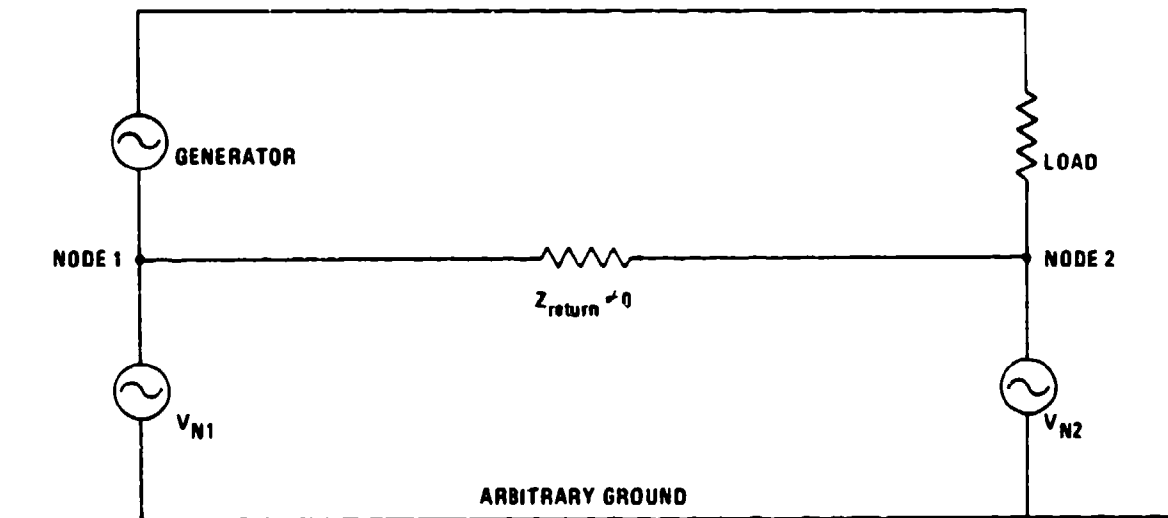


Figure 6-2. Energy Transfer Loop With Noise Sources in Ground System

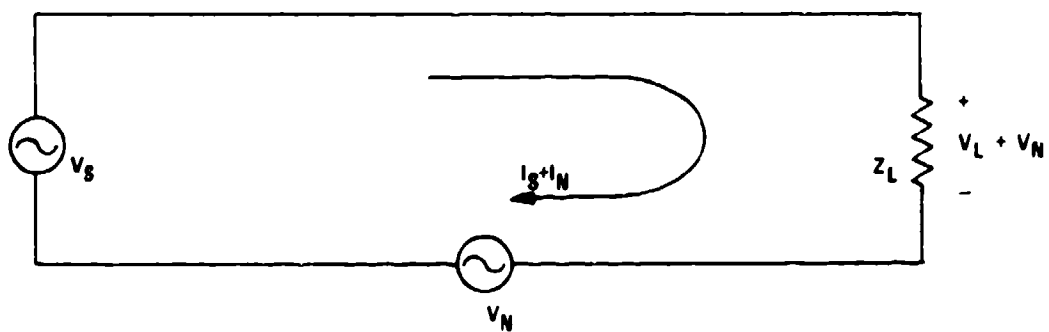
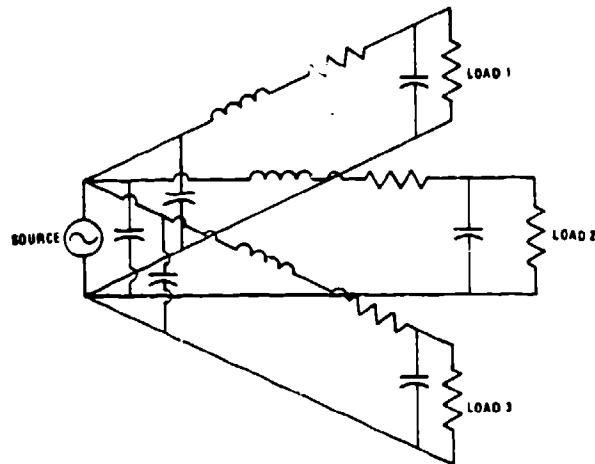
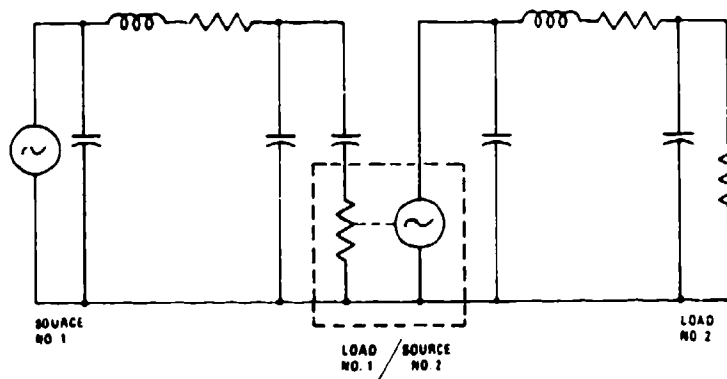


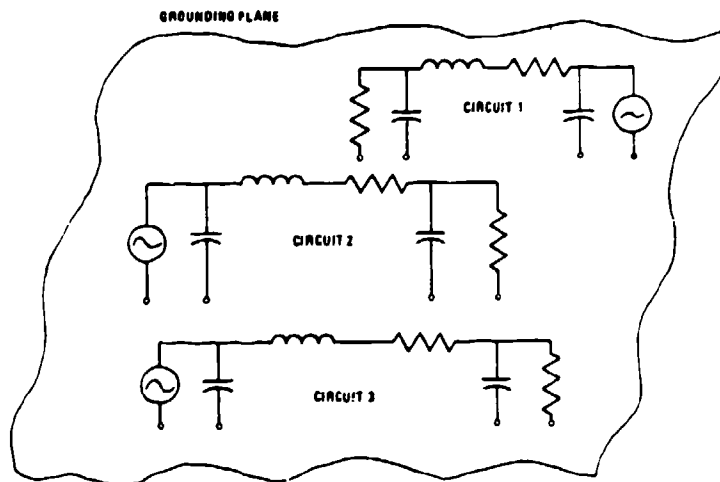
Figure 6-3. Equivalent Circuit of Non-Ideal Energy Transfer Loop



(a) ONE SOURCE DRIVING MULTIPLE LOADS.



(b) CASCADED PAIRS



(c) SEPARATE PAIRS CONNECTED TO COMMON GROUND PLANE.

Figure 6-4. Practical Combinations of Source-Load Pairs

At the equipment level, the individual signal reference planes for the various networks must be connected together to prevent personnel shock hazards (see Chapters 5 and 9) and to provide as near as possible, the same signal reference for all networks. Thus, the signal reference plane may extend over large distances within a facility. The assumption that this large reference plane provides zero impedance paths is not valid; the series inductance and resistance of the conductors forming the signal reference plane and the shunt capacitance to nearby conductive objects must be considered. Currents flowing in the signal reference plane will develop voltages across this impedance and will produce electric and magnetic fields around the conductors.

## 6.2 COUPLING MECHANISMS.

Coupling is defined as the means by which a magnetic or electric field produced by one circuit induces a voltage or current in another circuit. Interference coupling is the stray or unintentional coupling between circuits which produces an error in the response of one of the circuits. The possible sources of spurious signals and the mechanisms by which this interference is coupled into a susceptible circuit must be understood in order to guard against interference pickup by sensitive signal circuits.

The techniques for reducing pickup depend on the type of interference present. Interference is broadly classified by its coupling means; i.e., as either being conductive or free-space. Conductive coupling occurs when the interfering and the interfered-with circuits are physically connected with a conductor and share a common impedance. Free-space coupling occurs when a circuit or source generates an electromagnetic field that is either radiated and then received by a susceptible circuit or that is inductively or capacitively coupled (near-field) to a susceptible circuit.

### 6.2.1 Conductive Coupling.

Power lines entering a facility provide good conductive coupling paths from interference sources external to the facility. This interference can easily be conducted into a particular unit or piece of equipment via the power lines entering the equipment. Also, interference can conductively couple between various circuits inside the equipment on the common dc power lines. If one dc power supply is utilized with several circuits operating over various signal voltage and frequency ranges, the operation of one circuit may adversely affect the operation of other circuits. For example, if both the preamplifier and the power amplifier sections of an audio amplifier are supplied from a single dc power supply, variations in the relatively large current drawn by the power amplifier can appear as supply voltage variations at the preamplifier. These variations can be large compared to the operating signal levels in the preamplifier; the unwanted variations are amplified along with the desired signals and may produce distortion in the output of the amplifier.

Another set of paths for conductive coupling of interference is offered by the signal lines. In general, signal lines enter each facility and each unit or piece of equipment; i.e., such signal lines are usually necessary for interfacing electronic circuits. Interference can be conductively coupled into facilities, equipment, and circuits as readily by signal lines as by power lines.



The signal reference plane is another potential coupling path for unwanted signals between equipments and/or circuits. Since practical signal reference planes do not exhibit a zero impedance, any currents flowing in a signal reference plane will produce potential differences (voltages) between various points on the reference plane. Interfacing circuits referenced to these various points can experience conductively coupled interference in the manner illustrated by Figure 6-5. The signal current  $I_1$ , flowing in Circuit 1 of Figure 6-5 returns to its source through signal reference impedance  $Z_R$  producing a voltage drop  $E_{N1}$  in the reference plane. The impedance  $Z_R$  is common to Circuit 2, hence  $E_{N1}$  appears in Circuit 2 as a voltage in series with the desired signal voltage source,  $E_{S2}$ . This undesired source produces an interference voltage,  $V_{N2}$ , across the load of Circuit 2; similarly, the desired current,  $I_2$ , in Circuit 2 will produce interference in Circuit 1.

In a facility, the conductive coupling of interference through the signal reference plane of interfaced equipment can occur in a manner similar to that described above for internal circuitry. If Circuit 1 in Figure 6-5 represents two interfaced equipments and if Circuit 2 represents a different pair of interfaced equipments, then a current flowing in either circuit can produce interference in the other circuit as described. Even if the pairs of equipments do not use the signal reference plane as the signal return, the signal reference plane can still be the cause of coupling between equipments. Figure 6-6 illustrates the effect of a stray current,  $I_R$ , flowing in the reference (or ground) plane.  $I_R$  may be the result of the direct coupling of another pair of equipments to the signal reference plane, or it may be the result of free-space coupling to the signal reference plane. In either case,  $I_R$  produces a voltage  $E_N$  in the reference plane impedance,  $Z_R$ . This voltage produces a current in the interconnecting loop which in turn develops a voltage across  $Z_L$  in Equipment B. Thus, it is evident that interference can conductively couple via the signal reference plane to all circuits and equipments connected across the non-zero impedance elements of that reference plane.

#### 6.2.2 Free-Space Coupling.

Free-space coupling is the transfer of electromagnetic energy between two or more circuits not directly interconnected with a conductor. Depending on the distance between the circuits, the coupling is usually defined as either near-field or far-field. Near-field coupling can be subdivided into inductive and capacitive coupling, according to the nature of the electromagnetic field. In inductive coupling, a magnetic field linking the susceptible device or circuit is set up by the interference source or circuit. Capacitive coupling is produced by an electric field between the interference source and the susceptible circuit.

Radiation of energy by electromagnetic waves is the principle coupling mechanism in far-field coupling. The term "radiated coupling" is sometimes used to describe both near-field (inductive and capacitive) coupling and far-field coupling. However, radiated coupling is generally accepted to mean the transfer of energy from a source to a susceptible circuit by means of electromagnetic waves propagating through space according to the laws of wave propagation.

##### 6.2.2.1 Near-Field Coupling.

When two or more wires or other conductors are located near each other, currents and voltages on one wire will be inductively and capacitively coupled to the other wires. The wire acting as the interference source for this near-field coupling may be any conductor such as a high level signal line, an ac power line, a control line, or even a lightning down conductor. The currents or voltages induced into the other wires can further be conductively coupled into susceptible circuits.

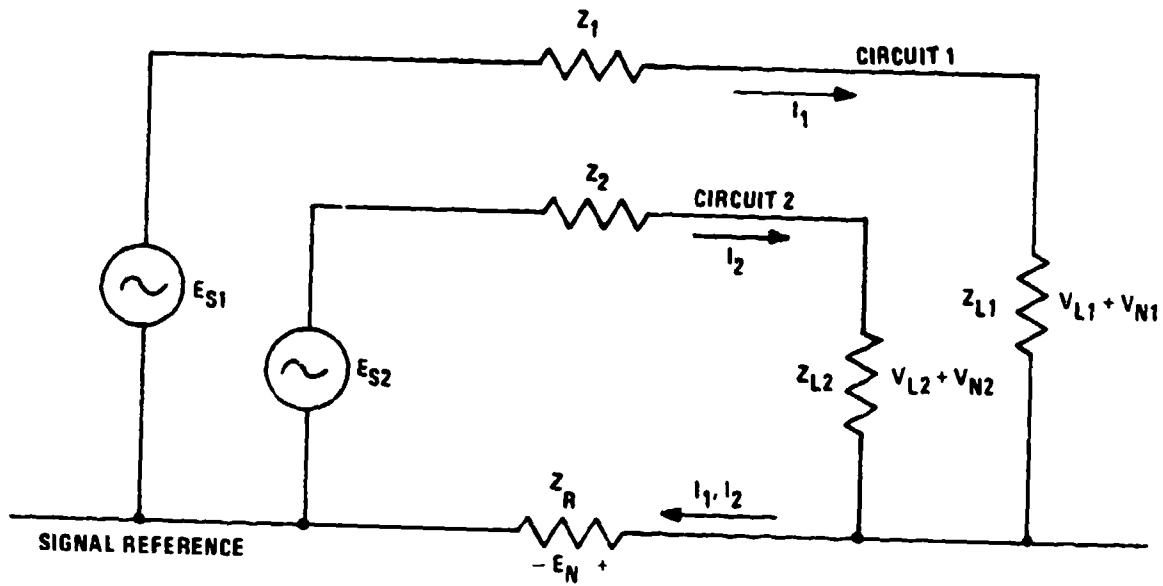


Figure 6-5. Coupling Between Circuits Caused by Common Return Path Impedance

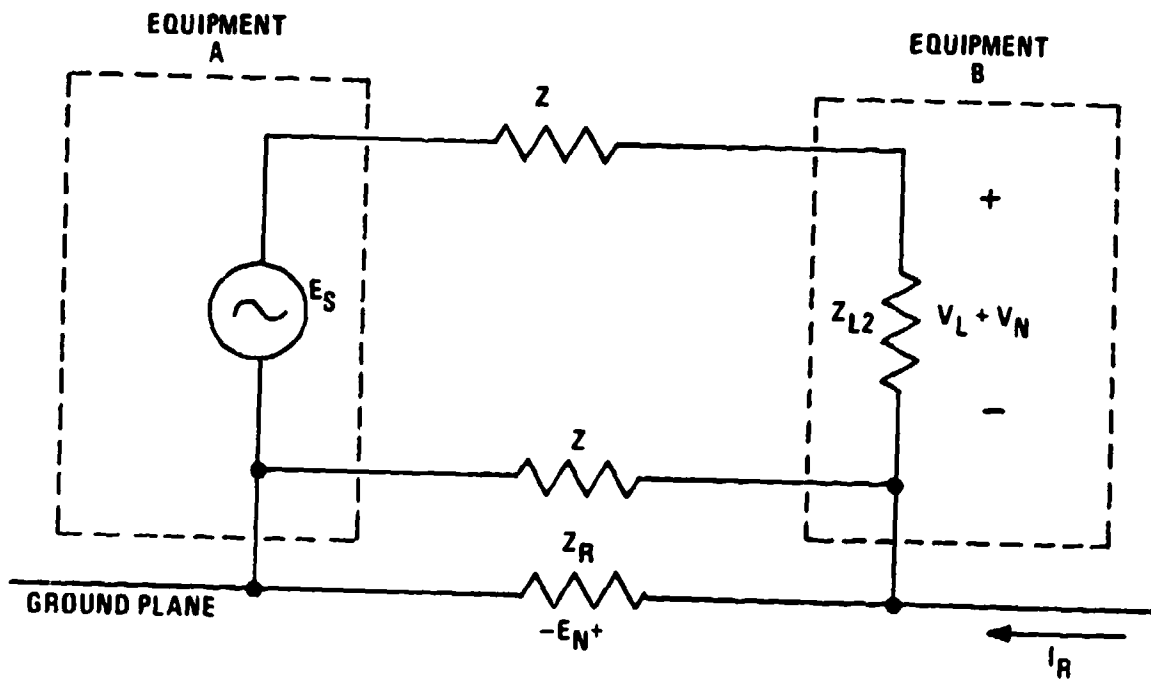


Figure 6-6. Conductive Coupling of Extraneous Noise into Equipment Interconnecting Cables

**6.2.2.2 Inductive Coupling.** The magnetic field surrounding a long, straight, current-carrying wire as shown in Figure 6-7 is the means for inductive coupling. This field can be determined from Ampere's law (6-1):

$$\oint \underline{H} \cdot d\underline{\ell} = i(t), \quad (6-1)$$

where  $\underline{H}$  is the magnetic field strength and  $d\underline{\ell}$  is a small element of length along the path of integration (any closed loop around the current  $i(t)$ ). Choosing a circle of radius  $r$  for the integration path in Equation 6-1 allows one to derive an expression for the magnetic field:

$$H = \frac{i(t)}{2\pi r} \text{ Amperes/meter.} \quad (6-2)$$

Thus the magnetic field strength surrounding a long straight wire carrying a current  $i$  is inversely proportional to the distance  $r$  from the wire, i.e.,  $H$  decreases as the distance from the wire increases.

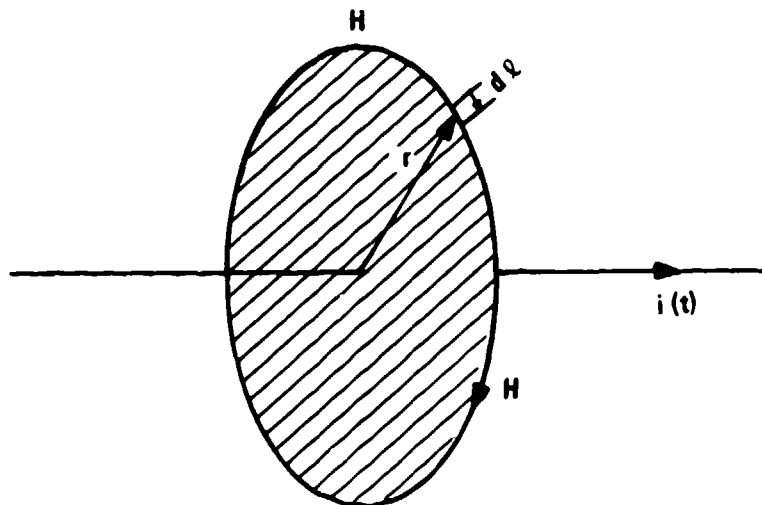


Figure 6-7. Magnetic Field Surrounding a Current-Carrying Conductor

This magnetic field will induce a voltage into a nearby signal circuit loop as illustrated in Figure 6-8. According to Faraday's law (6-1), the induced voltage is

$$v_i(t) = - \frac{d\psi}{dt} \quad \text{volts,} \quad (6-3)$$

where  $v_i(t)$  is the induced voltage and  $\psi$  is the total magnetic flux linking the susceptible circuit loop. This magnetic flux is given by

$$\begin{aligned} \psi &= \int_S \underline{B} \cdot d\underline{s}, \\ &= \mu \int_S \underline{H} \cdot d\underline{s} \quad \text{webers,} \end{aligned} \quad (6-4)$$

where  $\underline{B} = \mu \underline{H}$  is the magnetic flux density,  $\mu$  is the absolute permeability of the medium, and  $d\underline{s}$  is a small element of the loop area. Substituting Equation 6-2 into Equation 6-4 and integrating over the area of the loop in Figure 6-8 gives

$$\psi = \frac{\mu \ell i(t)}{2\pi} \ln \left( \frac{r_2}{r_1} \right) \quad \text{webers,} \quad (6-5)$$

where  $r_1$  and  $r_2$  are the distances from  $i(t)$  to the two sides of the loop which are parallel to  $i(t)$ , and  $\ell$  is the length of each of these sides (in meters). This equation gives the total magnetic flux linking a susceptible

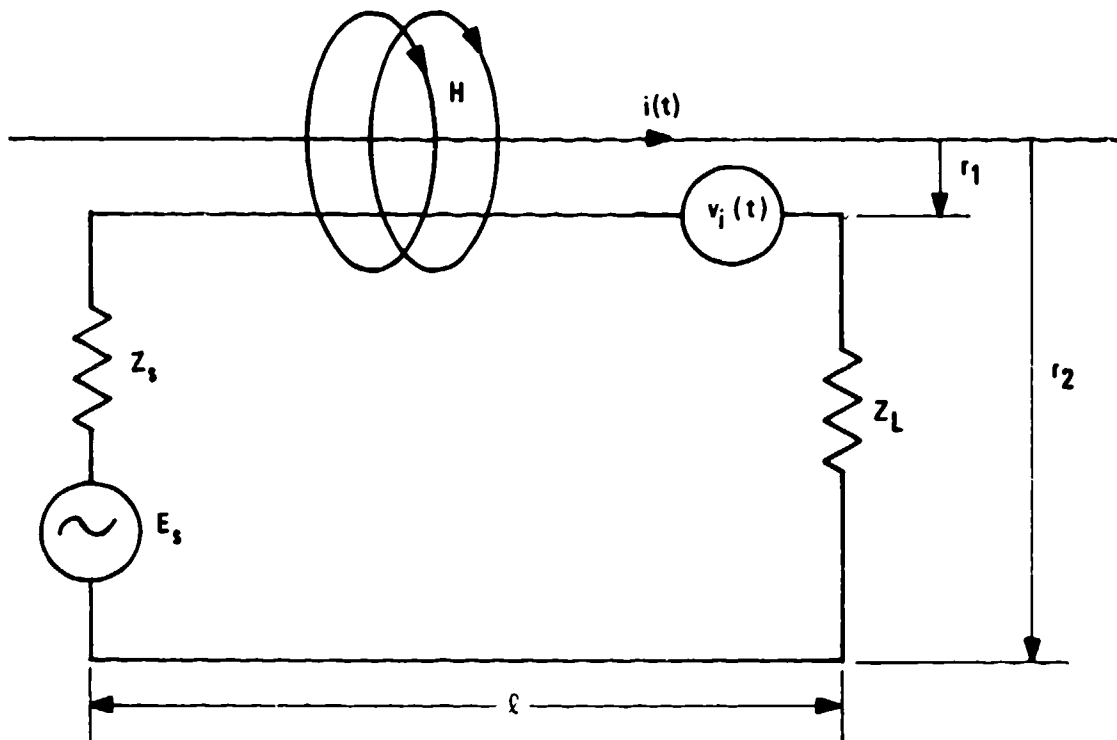


Figure 6-8. Illustration of Inductive Coupling

circuit loop in terms of the current flowing in a nearby conductor parallel to the sides of the loop. Substituting the total magnetic flux from Equation 6-5 into Equation 6-3 gives the voltage  $v_i(t)$ , induced in the circuit loop:

$$v_i(t) = \frac{-\mu l}{2\pi} \frac{di}{dt} \ln \left( \frac{r_2}{r_1} \right) \text{ volts.} \quad (6-6)$$

In free space,  $\mu = 4\pi \times 10^{-7}$  henrys/meter, and Equation 6-6 reduces to

$$v_i(t) = -2 \times 10^{-7} l \frac{di}{dt} \ln \left( \frac{r_2}{r_1} \right) \text{ volts.} \quad (6-7)$$

If the interference current in the nearby conductor is sinusoidal, i.e.,

$$i(t) = I_m \cos(\omega t + \theta), \quad (6-8)$$

then the maximum value of  $di/dt$  is

$$\left| \frac{di}{dt} \right|_{\max} = \omega I_m = 2\pi f I_m, \quad (6-9)$$

and the maximum value of the induced voltage in Equation 6-6 is

$$\left| v_i \right|_{\max} = \mu l f I_m \ln \frac{r_2}{r_1}, \quad (6-10)$$

where  $f$  is the frequency of  $i(t)$ .

In Equations 6-6 and 6-7, the induced voltage  $v_i(t)$  in a circuit loop with sides parallel to a current  $i(t)$  is expressed in terms of the dimensions of the loop, the distance of the loop from the current  $i(t)$  and time rate of change of  $i(t)$ . As can be seen from these equations, the induced voltage in a susceptible circuit loop increases with an increase in the loop area (either an increase in  $l$  or  $r_2$  or both), the frequency,  $f$ , or the time rate of change,  $di/dt$ , of the interfering sources, and increases with a decrease in the distance,  $r_1$ , between the interfering source and the loop.

The preceding equations indicate that the induced voltage is independent of the impedance of the susceptible circuit loop; i.e., the amplitude of the induced voltage is the same in a high impedance circuit as it is in a low impedance circuit. The desired signal voltages in low impedance circuits, however, are generally much lower than in high impedance circuits. Therefore, in low impedance circuits the induced voltage can be high relative to the signal voltage and thus more likely to produce significant interference in the circuit load. In high impedance circuits the same induced voltage may be small compared to the circuit signal voltages and thus not create any problems. For these reasons, low impedance circuits are usually more susceptible to inductive coupling than are high impedance circuits.

**6.2.2.3 Capacitive Coupling.** When signal conductors of two circuits are near each other as shown in Figure 6-9, a capacitance,  $C_c$  exists between the conductors. The value of this capacitance is a function of the geometry of the signal lines. For parallel wires,  $C_c$  is given by (6-2).

$$C_c = \frac{1.21}{\log\left(\frac{2r_c}{d}\right)} \times 10^{-11} \text{ farads/meter,} \quad (6-11)$$

where  $r_c$  is the distance between the two lines and  $d$  is the diameter of the wires. In a similar manner, a capacitance exists between each signal line and its return. If the signal line is parallel to its return, these capacitances can be calculated using Equation 6-11 by replacing  $r_c$  with  $r_1$  and with  $r_2$  (see Figure 6-9).

The interference source voltage,  $v_s(t)$  produces a current flow through the mutual capacitance,  $C_c$ , between the two signal conductors and develops an induced voltage,  $v_i(t)$ , in the susceptible circuit. The equivalent circuit for Figure 6-9 is given in Figure 6-10(a) where the parallel combination of  $Z_{S2}$  and  $Z_{L2}$  has been replaced by the equivalent impedance

$$Z_2 = \frac{Z_{S2}Z_{L2}}{Z_{S2} + Z_{L2}}, \quad (6-12)$$

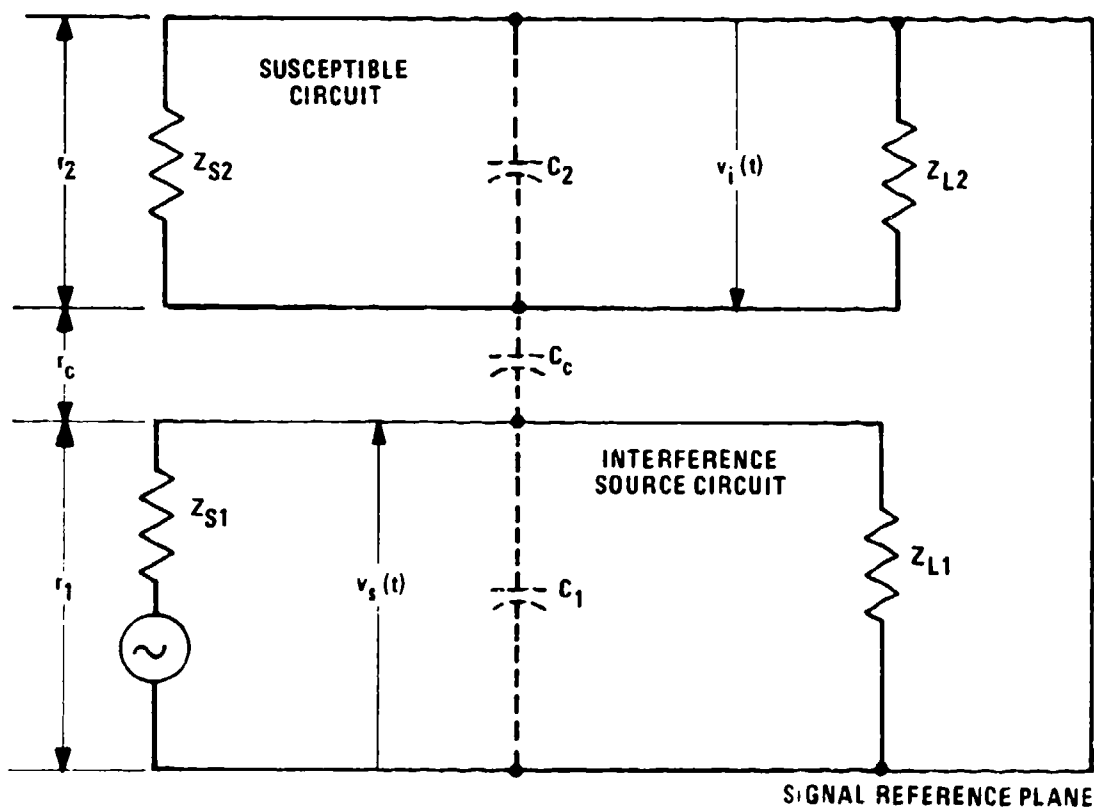


Figure 6-9. Illustration of Capacitive Coupling

and time-varying voltages have been replaced by their ac steady state phasors. The induced voltage (ac steady state assumed) in the susceptible circuit is

$$V_i = \frac{Z_{C2}Z_2/(Z_{C2} + Z_2)}{Z_{CC} + Z_{C2}Z_2/(Z_{C2} + Z_2)} V_s, \quad (6-13)$$

where

$$Z_{C2} = \frac{1}{j\omega C_2} \quad (6-14)$$

and

$$Z_{CC} = \frac{1}{j\omega C_c} \quad (6-15)$$

Substitution of Equations 6-14 and 6-15 into Equation 6-13 yields:

$$V_i = \frac{j\omega Z_2 C_c}{1 + j\omega Z_2 (C_2 + C_c)} V_s \quad (6-16)$$

where  $\omega = 2\pi f$ . If  $j\omega Z_2 C_c \ll 1$ , which is generally true at lower frequencies, the equivalent circuit of Figure 6-10(b) is applicable and

$$V_i = j2\pi f Z_2 C_c V_s \quad (6-17)$$

At higher frequencies, the equivalent circuit of Figure 6-10(c) is applicable and

$$V_i = \frac{C_c}{C_2 + C_c} V_s \quad (6-18)$$

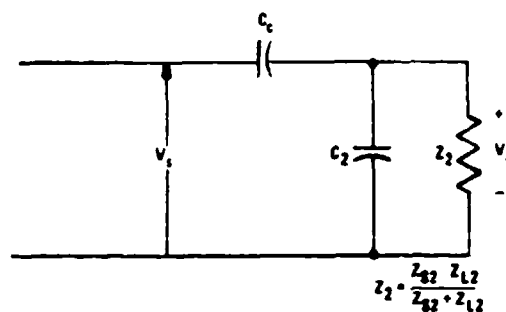
These equations illustrate the induced voltage,  $V_i$ , which is capacitively coupled into a susceptible signal circuit from a nearby signal conductor, is dependent on the amplitude and frequency of the interference source voltage,  $V_s$ , the values of the coupling capacitance,  $C_c$ , the stray capacitance in the susceptible circuit,  $C_2$ , and on the magnitude of the impedance of the susceptible circuit. At low frequencies, Equation 6-17 indicates that the induced voltage increases with either an increase in the coupling capacitance or an increase in the impedance of the susceptible loop. Similarly, at high frequencies the induced voltage as given in Equation 6-18 increases with either an increase in the coupling capacitance or a decrease in the stray capacitance of the susceptible circuit. It should also be noted that the value of the interference source voltage,  $V_s$ , depends upon the stray capacitance in the interference source circuit,  $C_1$  in Figure 6-9.

If the impedance of the susceptible circuit is totally resistive, i.e., if  $Z_2 = R_2$ , the induced voltage is given by Equation 6-16 as

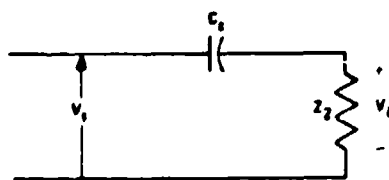
$$V_I = \frac{j2\pi f R_2 C_c}{1 + j2\pi f R_2 (C_2 + C_c)} V_s, \quad (6-19)$$

and the magnitude of the ratio of the induced voltage to the interference (sinusoidal) voltage is

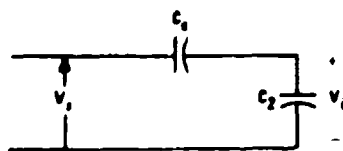
$$\left| \frac{V_I}{V_s} \right| = \sqrt{\frac{(2\pi f R_2 C_c)^2}{1 + (2\pi f R_2 (C_2 + C_c))^2}}$$



(a) TRUE EQUIVALENT CIRCUIT



(b) LOW FREQUENCY APPROXIMATION



(c) HIGH FREQUENCY APPROXIMATION

Figure 6-10. Equivalent Circuit of Network in Figure 6-9



$$= \frac{C_c}{C_2 + C_c} \sqrt{\frac{R_2^2}{R_2^2 + \left[ \frac{1}{2\pi f(C_2 + C_c)} \right]^2}} \quad (6-20)$$

This ratio increases almost linearly with  $R_2$  until  $R_2$  approaches the value  $\frac{1}{2\pi f(C_2 + C_c)}$ ; i.e., the reactance of  $C_2$  and  $C_c$  in parallel. For larger values of  $R_2$ , the ratio asymptotically approaches  $C_c/(C_2 + C_c)$ . The behavior of this voltage ratio with frequency is illustrated in Figure 6-11. The ratio is zero at dc and asymptotically approaches  $C_c/(C_2 + C_c)$  as the frequency is increased. Equation 6-20 and Figure 6-11 illustrate again that the voltage capacitively coupled into the susceptible circuit increases with an increase in the total resistance of the circuit and with an increase in frequency. Resonances can occur and change the amount of capacitive coupling if the impedance of the susceptible circuit contains inductive reactance, but such resonances usually only produce noticeable effects at higher frequencies.

#### 6.2.2.4 Far-Field Coupling.

Radiation is the means by which energy escapes from a conductor and propagates into space. The conductor does not have to be specifically designed to radiate energy; it may be any current carrying conductor, e.g., a signal line, a power line, or even a ground lead.

Algebraic expressions for the electromagnetic fields surrounding a current carrying conductor are usually expressed as the sum of three terms. Each term is inversely proportional to a power of the distance,  $r$ , from the conductor, i.e., each term is proportional to either  $1/r$ ,  $1/r^2$ , or  $1/r^3$ . Close to the conductor (near field), the  $1/r^2$  and  $1/r^3$  components dominate and the electromagnetic energy oscillates between the space surrounding the conductor and the conductor itself; zero average energy is propagated by the near field terms.

Outside the near field region, the  $1/r$  term predominates. In this far field region, radiated energy that has escaped is propagating away from the "antenna" through space. The mechanism of energy radiation can be visualized (6-3) by considering the finite time required for the electromagnetic fields to propagate between two points in space. Current flows through an antenna at the frequency of the applied signal, and the polarity of the field produced by this current is reversed at this same frequency. When a positive charge is present at one end of the antenna, an equal but, negative charge is present at the other end and an electric field in the vicinity of the antenna will be established between the charges. As the current changes direction, the charges will reverse positions; the electromagnetic field will collapse and be re-established in the opposite direction. If the frequency of the applied signal is low, sufficient time will exist between reversals for practically all the energy stored in the field to be returned to the circuit and very little radiation will occur. If, however, the frequency is high and the charges reverse quickly, a field in the opposite direction is formed near the wire before a substantial amount of the field energy can return to the circuit. This part of the field is thus separated from the antenna and propagates outward through space as an electromagnetic wave.

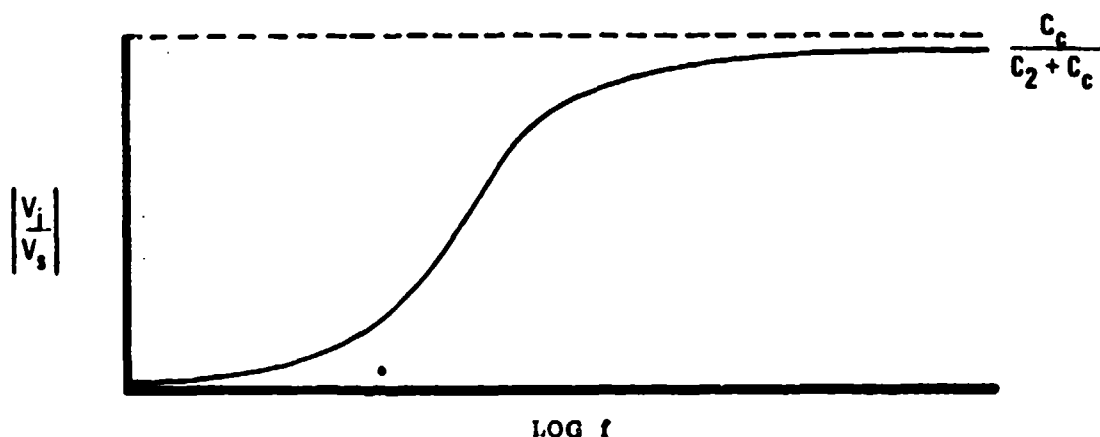


Figure 6-11. Characteristic Voltage Transfer Curve for Capacitive Coupling

This method of visualizing radiation from a wire or antenna is illustrated for a dipole antenna in Figure 6-12. Figure 6-12(a) shows a dipole when the charges are maximum at the ends of the antenna. As the current flow reverses directions and the charges move back toward the center, the electric field lines collapse as in (b). Since the field moves with a finite velocity, there is not enough time for all the field lines to return to the antenna. When the ends of these lines meet at the center of the antenna and the charge on the antenna is zero, the field lines that have not collapsed will close on themselves and continue to exist as closed loops as illustrated in (c). The antenna charges move in the opposite direction as shown in (d), and the oppositely directed electric field pushes away the previously detached loop as shown in (e). This procedure continues with the fields in the opposite direction, and a cycle is completed when the fields near the antenna return to their original state. These cycles repeat at the frequency of the applied signal, and an electromagnetic field propagates outward from the antenna at the speed of light. Although only the electric field is illustrated, there is an associated magnetic field in accordance with Maxwell's equations (6-1). The magnetic field consists of concentric circles surrounding the antenna and expanding radially as the electric field propagates outward. These outward propagating electric and magnetic fields represent energy flowing away from the antenna. Therefore, the antenna radiates energy into the surrounding space.

In a reciprocal manner, wires and conductors located in a radiated field have currents induced in them and act as receiving antennas for incident electromagnetic energy. These induced currents in the wires can then be conducted into associated signal circuitry as interference (see Section 6.2.1). The amplitude of the resulting interference depends on the strength of the electromagnetic field in the vicinity of the wire and on the efficiency of the wire as an antenna. The strength of this field is a function of the distance from the radiating wire, the efficiency of the radiating wire as an antenna, and the amplitude and frequency of the signal on the radiating wire. The efficiency of a wire or other conductor as either a receiving or a radiating antenna is a function of the length of the wire relative to the wavelength of the signal.

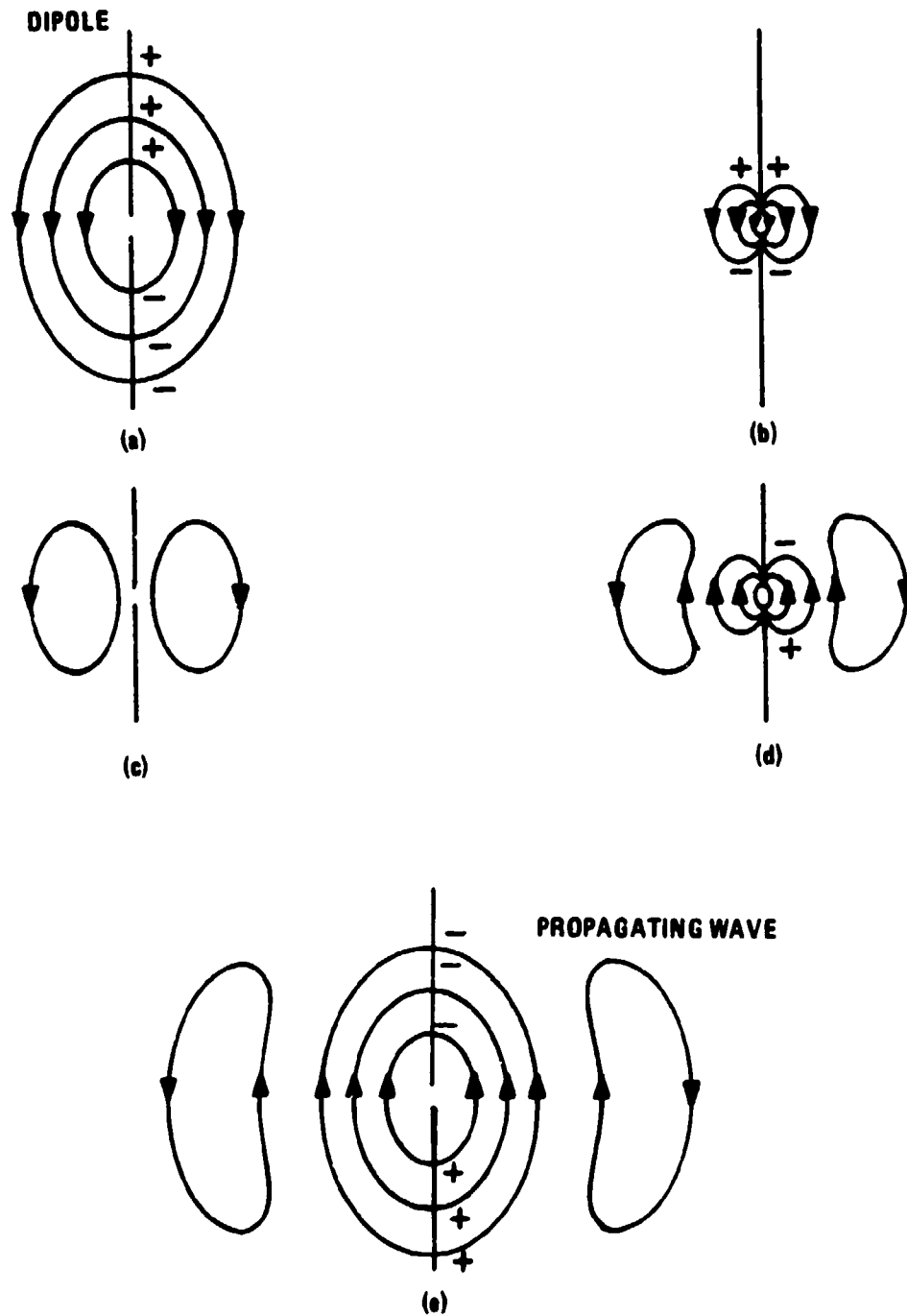


Figure 6-12. Electric Field Patterns in the Vicinity of a Radiating Dipole

One way of evaluating the efficiency of a wire as an antenna is to compare its radiation resistance with the radiation resistance of a quarter-wave ( $\lambda/4$ ) antenna. The radiation resistance of an antenna is the resistance which would consume the same amount of power as is radiated by the antenna. Thus the radiation resistance is a direct measure of the energy radiated from the antenna. A monopole antenna one-quarter of a wavelength long has a radiation resistance of 36.5 ohms (6-4). An antenna which transmits or receives ten percent or less of the energy that would be transmitted or received by a  $\lambda/4$  monopole can be considered relatively inefficient. Thus an inefficient antenna would exhibit a radiation resistance of 3.65 ohms or less. Monopoles of length  $\lambda/11$  meet this criterion (6-4). Greater convenience in calculations results if  $\lambda/10$  is chosen instead of  $\lambda/11$ . Thus  $\lambda/10$  is chosen to represent the length below which a conductor does not perform effectively as an antenna.

### 6.3 COMMON-MODE NOISE.

Common-mode noise is an unwanted noise voltage which appears identically on both sides of a signal line when measured from the system ground or common point. It, like normal-mode noise, can be caused by resistive coupling, capacitive coupling, or magnetic coupling from the unwanted source. In addition, many measuring transducers intentionally have a dc or ac common-mode voltage present on both output lines, the presence of which is necessary for proper operation of the transducer. Although not a noise source, these common-mode voltages require careful design and use of data and instrumentation amplifiers to prevent interference with the desired signal components.

The source of most common-mode noise is resistive coupling between separate ground points in a circuit or system. A simple example of this is illustrated in Figure 6-13. An oscilloscope probe is used to couple a signal from some point in a circuit to the oscilloscope terminals. The probe ground is connected to circuit ground which is in turn referenced through the facility ground system. Since there are generally currents flowing in the facility ground system (these are primarily at the 60 Hz power line frequency), it follows that the ground reference potential for the circuit is different from that for the oscilloscope. This difference in potential is produced by the flow of the stray ground currents through the impedance of the facility ground system. Thus, both the ground reference for the circuit and the signal point in the circuit have identical noise voltages impressed on them with respect to the ground reference for the oscilloscope. This noise is called common-mode noise by virtue of the fact that is common to all points in the circuit, including the circuit ground. Not only do these noise sources introduce measurement errors but they also produce interference between interconnected equipments.

Resistively coupled common-mode noise can also occur in a single equipment rather than between equipments. The coupling arises from multiple signal currents and power frequency currents flowing in a common ground lead, chassis, or ground plane.

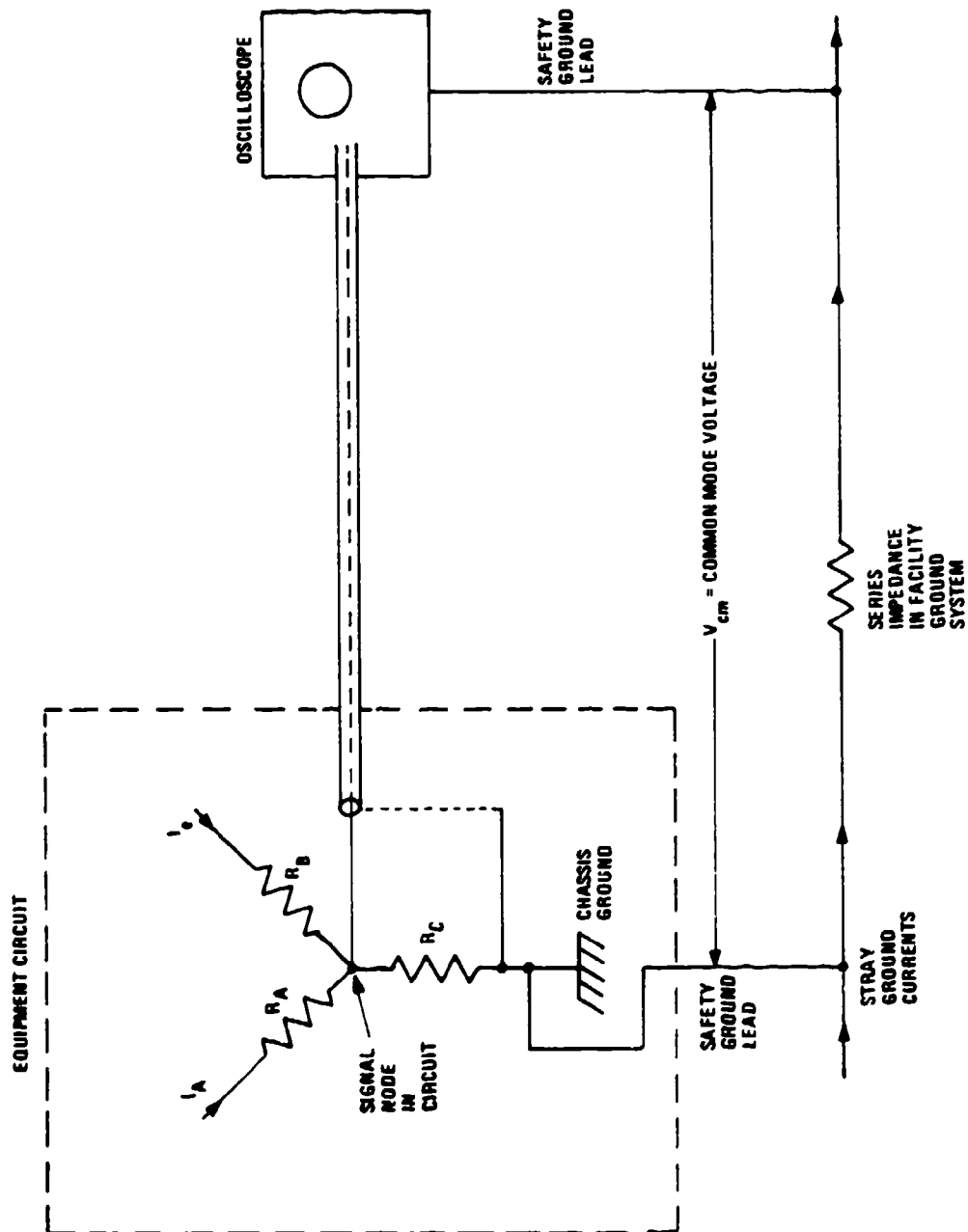


Figure 6-13. Illustration of Conductively-Coupled Common-Mode Noise

### 6.3.1 Basic Theory of Common-Mode Coupling.

The mechanism of common-mode coupling can be explained with reference to Figure 6-14. In this figure,  $V_s$  represents some signal voltage from an unbalanced source, i.e., the output signal of some transducer or measuring amplifier, and  $R_s$  is the output impedance of this source. The source is connected to the input terminals of some electronic device which is modeled as a two-terminal pair amplifier in the figure.  $R_1$  and  $R_2$  are the series resistances in the interconnecting cables between the source and amplifiers. The voltage source  $V_{cm}$  with output resistance  $R_{cm}$  represents a common-mode noise voltage source which causes the signal source to be at some voltage when measured with respect to the ground reference of the amplifier output. In Figure 6-14, the impedances  $Z_1$  and  $Z_2$  represent the input impedances of the two amplifier terminals. In a differential amplifier, these impedances are normally very high, however, in a single ended amplifier, one is high and the other is very low since it is tied directly to the ground reference terminal.

The analysis of the circuit in Figure 6-14 is complicated enough to make it difficult to reach conclusions without excessive algebra. Normally,  $R_{cm}$  is small and can be neglected. With this approximation, it can be shown that the output voltage of the amplifier is given by

$$\begin{aligned}
 V_o &= KV_x \\
 &= K \left( \frac{Z_1}{R_s + R_1 + Z_1} - \frac{Z_2}{R_2 + Z_2} \right) V_{cm} \\
 &\quad + \left( \frac{KZ_1}{R_s + R_1 + Z_1} \right) V_s,
 \end{aligned} \tag{6-21}$$

where  $K$  is the voltage gain of the amplifier.

There are two signal contributions to the output signal  $V_o$  in Equation 6-21: the desired signal and the undesired common-mode noise. There are three ways in which the common-mode noise term can be reduced. These are as follows:

- a. Decrease  $V_{cm}$  - By decreasing  $V_{cm}$ , the common-mode noise voltage at the output terminals decreases proportionally.

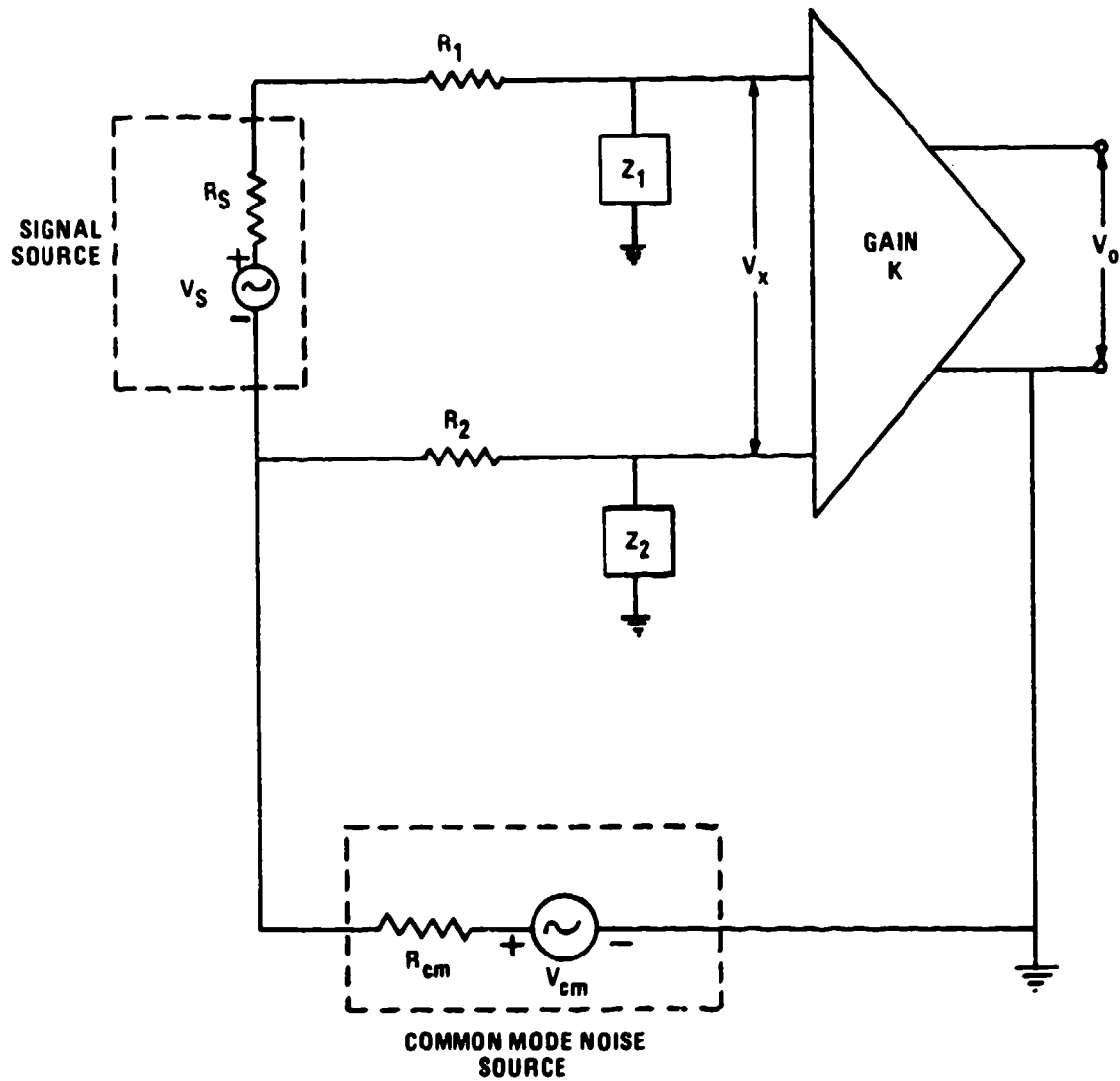


Figure 6-14. Common-Mode Noise in Unbalanced Systems

- b. Balance the Two Amplifier Inputs - If  $R_1$  and  $R_2$  are manipulated such that

$$\frac{Z_1}{R_s + R_1 + Z_1} = \frac{Z_2}{R_2 + Z_2}, \quad (6-22)$$

the common-mode noise voltage at the amplifier output terminals can be made to vanish.

- c. Increase  $Z_1$  and  $Z_2$  - If  $Z_1$  is sufficiently large compared to  $R_s + R_1$ , and  $Z_2$  is sufficiently large compared to  $R_2$ , then the common-mode noise voltage at the amplifier output terminals will be diminished. This approach normally requires a differential amplifier with carefully shielded input signal lines.

In the case of balanced signal sources or transducers, the basic circuit and equations differ from those given in Figure 6-14 and by Equation 6-21. Figure 6-15 shows a balanced source with an output voltage  $V_s$  and output resistance  $R_s$  connected to the two inputs of an amplifier. In this case, the center tap of the source is connected to the ground reference terminal. As before, if it is assumed that  $R_{cm}$  is small, it can be shown that  $V_o$  is given by

$$\begin{aligned} V_o &= KV_x, \\ &= \left( \frac{Z_1}{\frac{R_s}{2} + R_1 + Z_1} - \frac{Z_2}{\frac{R_s}{2} + R_2 + Z_2} \right) V_{cm} \\ &\quad + K \left( \frac{Z_1}{\frac{R_s}{2} + R_1 + Z_1} + \frac{Z_2}{\frac{R_s}{2} + R_2 + Z_2} \right) \frac{V_s}{2} \end{aligned} \quad (6-23)$$

The same conclusions regarding the minimization of the common-mode noise component at the amplifier output apply in this circuit as for the unbalanced source. However, in this case the amplifier must have a differential input stage. Otherwise, one-half of the source would be shorted out. In Figure 6-14, the amplifier can have single-ended or differential inputs.

The common-mode rejection (CMR) ratio of an amplifier is the gain of the amplifier ( $K$ ) multiplied by the common-mode noise voltage ( $V_{cm}$ ) and divided by the amplifier output due to  $V_{cm}$ . The CMR ratio describes a circuit's ability to avoid converting common-mode noise to normal-mode noise. Expressed as a positive quantity, the CMR ratio is given by

$$CMR = \left| \frac{KV_{cm}}{V_o} \right| \quad V_s = 0 \quad (6-24)$$



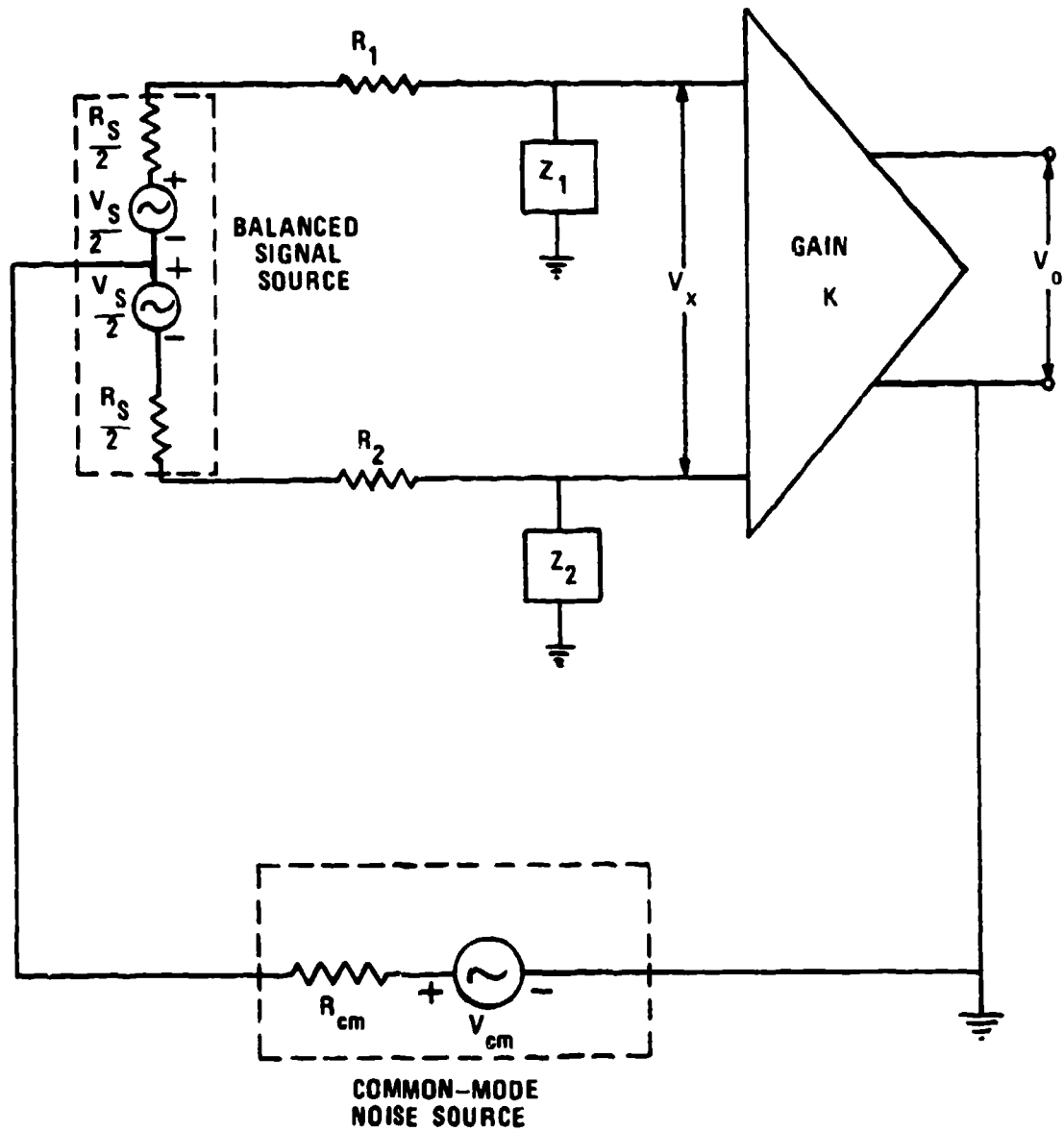


Figure 6-15. Common-Mode Noise in Balanced Systems

Ideally, the CMR of an amplifier should be infinite, or as large as possible. Under the worse case conditions, CMR = 1. As it is defined, the CMR conveys a measure of how well the amplifier can reject a common-mode noise signal at its input. Typical values for a good differential amplifier with balanced input impedances are in the vicinity of CMR = 1000. Often this is expressed in decibels which, in this case, would be CMR = 60 dB.

The CMR for the amplifier in Figure 6-14 is easily derived from Equation 6-21 to be

$$\text{CMR} = \left| \frac{1}{\frac{Z_1}{R_s + R_1 + Z_1} - \frac{Z_2}{R_2 + Z_2}} \right| \quad (6-25)$$

**6.3.2 Differential Amplifier.** A differential amplifier is designed to make  $Z_2$  large compared to  $R_2$  and  $Z_1$  large compared to  $R_1 + R_s$ . Since  $Z_1$  and  $Z_2$  are normally functions of frequency, it can be seen that the CMR will also be a function of frequency. Typically  $Z_1$  and  $Z_2$  are resistors shunted by capacitors. Thus, it can be seen that the CMR will inevitably decrease with increasing frequency when the capacitive reactance becomes smaller than the resistor. Consequently, a high CMR is difficult to achieve at high frequencies.

**6.4 MINIMIZATION TECHNIQUES.** Signal interaction, i.e., interference, can be minimized by reducing the coupling between the signal systems by modifying the signal systems in such a manner that interaction between the systems does not produce interference in either one, by eliminating the source of the interference, and by filtering the interference out of the susceptible signal system.

**6.4.1 Reduction of Coupling.** The techniques for reducing coupling include minimizing the impedance of the reference plane, increasing the spatial separation between the signal systems, shielding the systems from each other, reducing the loop area of each signal system, and balancing the signal lines in each system.

**6.4.1.1 Reference Plane Impedance Minimization.** Minimizing the impedance of the signal reference plane lowers the potential difference between any two points in the reference plane, thereby reducing the conductive coupling of interference in susceptible circuits referenced to these points. The impedance of the signal reference plane is reduced by minimizing both the resistance ( $R$ ) and the series reactance ( $X$ ) of the conductors forming the reference plane. The resistance decreases with a decrease in either the length of the conductors or the signal frequency (because of skin effect - see Section 5.2.2.1) and with an increase in conductor cross-sectional area. The reactance also decreases with a decrease in the signal frequency and with a decrease in the inductance of the conductors; the inductance is a function of both the conductor length and cross-sectional area. The impedance of the signal reference plane can be reduced by making the reference plane conductors as short as possible and by using conductors with cross-sectional areas as large as practical. The overall impedance of the signal reference plane also depends upon the establishment of low impedance bonds between ground conductors. (The various aspects of bonding and bond resistance are discussed in Chapter 7.)

**6.4.1.2 Spatial Separation.** Inductive or capacitive coupling can be reduced by increasing the physical distance between signal circuits. As can be seen from Equation 6-6 and Equations 6-11 and 6-16, increasing the separation between the interfering circuit and the susceptible circuit exponentially decreases the voltage coupled into the susceptible circuit.

**6.4.1.3 Reduction of Circuit Loop Area.** Reducing the loop area of either the interference source circuit or the susceptible circuit will decrease the inductive coupling between the circuits. Equation 6-6 shows that the inductively coupled voltage can be minimized by reducing the length ( $l$ ) or the width ( $r_2 - r_1$ ) of the susceptible circuit. This width can be minimized by running the signal return adjacent to the signal conductor and, hence, reducing the loop area of the susceptible circuit. A preferable approach is to twist the signal conductor with its return. The use of twisted wires reduces the inductively coupled voltages since the voltage induced in each small twist area is approximately equal and opposite to the voltage induced in the adjacent twist area.

**6.4.1.4 Shielding.** Another effective means for the reduction of coupling is the use of shields around the circuits and around interconnecting lines. Principles of shielding are presented in Chapter 8.

**6.4.1.5 Balanced Lines.**

In situations where signal circuits must be grounded at both the source and the load, and hence, establish conductive coupling paths, the use of balanced signal lines and circuits is an effective means of minimizing the conductively coupled interference. In a balanced circuit, the two signal conductors are symmetrical with respect to ground. At equivalent points on the two conductors the desired signal is opposite in polarity and equal in amplitude relative to ground. A common-mode voltage will be in phase and will exhibit equal amplitudes on each conductor and will tend to cancel at the load. The amount of cancellation depends upon the degree to which the two signal lines are balanced relative to ground.

If the signal lines are perfectly balanced, the cancellation would be complete and the coupled interference voltage at the load will be zero. If the source and load are not normally or cannot be operated in a balanced mode, balanced-to-unbalanced transformers or other coupling devices should be used at both the source and load ends of the signal line.

**6.4.2 Alternate Methods.**

Several alternate methods exist for minimizing interference besides the reduction of coupling. The first technique consists of actual circuit modification. For example, the signal frequency of either the interfering source or the susceptible circuit can be changed such that the signals do not interfere with one another. Similarly, the desired signal can be transposed to another frequency range or to a type of signal not affected by the noise. An example of the former is the conversion of the desired signal to VHF/UHF or microwave while an example of the latter is the use of acoustic coupling and electro-optical transmission. Through the use of one of these techniques, the frequency of transmission over that portion of the path susceptible to pickup is such

that the receiving and detection devices do not respond to the noise signal. As another option, the amplitude of the interference source or the sensitivity of the susceptible circuit can be decreased to reduce the interaction between the two circuits. Further, the type of modulation used in one or both circuits can be changed to minimize the interference.

Another technique is the elimination of the interference source. Although this may seem like a trivial solution, it is a valid alternative in many situations. For example, the source of interference may be a rusty joint which can be eliminated by proper bonding.

A third alternative is the use of filters. The majority of interfering signals, even if they are free-space coupled to the signal and power lines, are conductively coupled into the susceptible circuit. The proper application of filters to both the signal and power lines can reduce this coupling.

**6.5 FACILITY AND EQUIPMENT REQUIREMENTS.** The interference rejection principles identified in this chapter are responsible, in part, for many of the recommendations contained elsewhere in this volume and in Volume II. For example, intersite or interbuilding common-mode noise voltages in the earth contribute to the need for a low resistance of 10 ohms to earth at each facility. Even a resistance to earth of as low as 10 ohms may not, however, alleviate all common mode noise on a data cable connecting two separate locations or buildings. While a low resistance may help, there will always be potential differences between any two rods in the ground. The use of shielded, balanced twisted pair for all lower frequency equipment interfaces recommended in Volume II, is intended to provide additional common-mode rejection to those unavoidable noise voltages which exist in any facility. This is not to say that the sources of noise in a facility cannot be controlled. In fact, much can be done by equalizing the load between the phases of the ac distribution system; by insuring that the neutral is grounded only at the service disconnecting means as recommended in Volume II; and by limiting the quantity of leakage current from power line filter capacitors by using the smallest acceptable value of capacitance or by sharing common filtered lines with several pieces of equipment.

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**2.11.2.1 Electrode Resistance.** The resistance of ground,  $R$ , of a single vertical electrode of length  $l$  in cm, and radius  $\alpha$  in cm, emplaced in homogeneous soil of resistivity  $\rho$  (ohm - cm) is found from:

$$R \text{ (ohms)} = \frac{\rho}{2\pi l} \left[ \ln \frac{4l}{\alpha} - 1 \right] \quad (2-77)$$

This equation may be used to estimate the penetration depth of conductive salt solutions in the soil adjacent to the treated backfill. Since the backfill is conductive, the electrode radius therefore is not just that of the metallic electrode, but initially the diameter of the hole filled with treated backfill. This large composite electrode is referred to as the effective electrode. For a constant ground temperature, any reduction in electrode resistance of a frozen saturated soil with time should be related to an increase in effective electrode diameter, presumably through salt movement. This increase can be determined by the soil resistivity  $\rho$  from equation 2-77 using the resistance to ground of the test electrode and the effective electrode radius measured at the time of installation. Periodically, after installation, the resistance to ground should be remeasured and the effective electrode radius can be calculated using the following form of equation 2-77 and using the soil resistivity calculated earlier:

$$\alpha \text{ (cm)} = \frac{4l}{\exp \left[ 1 + (2\pi l R / \rho) \right]} \quad (2-78)$$

#### **2.11.2.2 Installation and Measurement Methods.**

**2.11.2.2.1 Electrode Installation.** Holes can be drilled with augers designed for use in frozen ground with hole diameters ranging from 3.8 cm (1-1/2 in.) to 91.4 cm (3 ft) and depths seldom greater than 2 m (8 ft). Hand-held equipment, consisting of an electric drive or a 5-hp gasoline-powered drill can also be used for most of the shallow, smaller-diameter holes. Both units could be used with a coring auger to drill holes up to 11 cm (4 in.) in diameter in fine-grained frozen soils. A truck-mounted auger can be used for the larger-diameter vertical holes drilled in coarse-grained materials. The horizontal electrodes can be hand-pushed and then driven into the thin seasonally thawed layer.

Military 6.8 kg (M2A3) shaped charges (used only by qualified personnel) can also be employed to produce vertical holes. Their similar performances in a range of frozen materials, with penetration approaching the length of standard electrodes, make this charge size ideal for electrode installation. The volume of several of the drilled holes can also be expanded by using C-4 block explosives.

**2.11.2.2.2 Backfill.** Reduction of contact potential is important in establishing a good electrical ground. In frozen soil, ice can form around the electrode, causing high contact resistance. Ice formation on the rod surface is likely since the rod is easily chilled by exposure of the upper end to low air temperatures. The beneficial effect of pouring untreated water around an electrode will only be short-term in cold environments. Therefore, the use of conductive backfill with a low freezing point becomes paramount to attain good ground or earth contact.

The backfill can be prepared by mixing salt and local soil or by saturating the soil backfill with a salt-water solution as shown in Figure 2-31. Backfill other than soil can also be used because soil is not always easily recovered from some drilled or blasted holes and because unfrozen material is difficult to find during the winter. Absorbent paper saturated with a salt solution and compacted in the hole around the electrode can also be used as a soil substitute.

The amount of salt added to the backfill is determined by preliminary laboratory conductivity measurements of several salt-soil mixtures. Salt may be added to silt and to a fine sand to obtain mixtures of from 0 to 20% salt based on the weight of the air-dried soil. Distilled water can be added to the salt-soil mixtures to obtain several soil moisture levels up to saturation for both materials. The soils should be compacted into a cylindrical plexiglass ring, which is clamped between electrodes for resistivity measurements at 1 kHz. Figure 2-32 shows the resistivity for two soils as a function of salt concentration at several volumetric moisture contents. A salt-soil mixture containing 1% salt results in a dramatic decrease in resistivity, with little effect after 5% salt for most moisture levels. Therefore, a 5% salt by weight is recommended for backfill as it produces a very conductive salt-soil mixture with the least amount of salt.

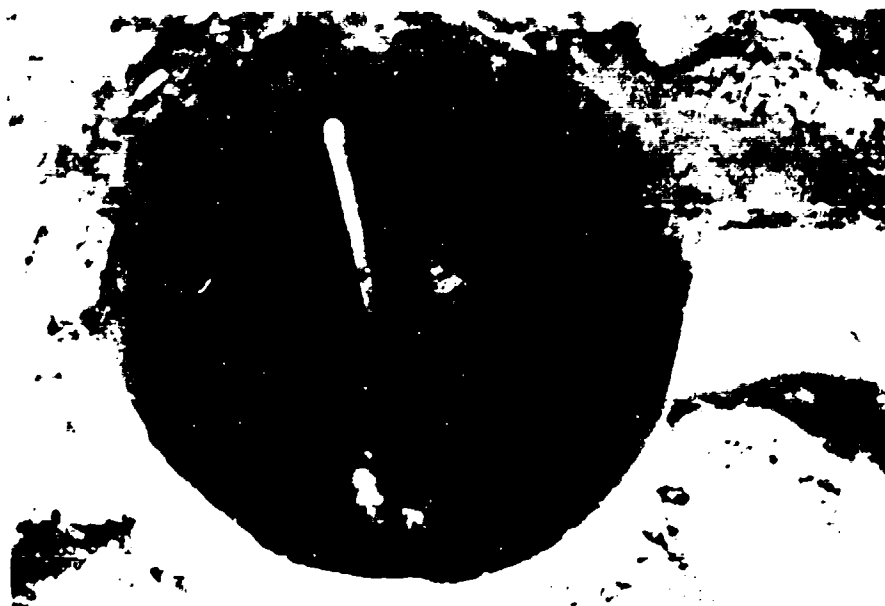


Figure 2-31. Installation of an Electrode During the Process of Backfilling with a Salt-Soil Mixture

Salt solution may also be poured around shallow-driven horizontal electrodes to minimize contact resistance during freezeback. These salt solutions in general may have concentrations on the order of 50-100%. Figure 2-33 shows a configuration of such horizontal electrodes placed in a thawed active layer.

Curves showing resistance-to-ground for metallic electrodes having various backfills are shown in Figures 2-34 through 2-38. Large seasonal variations are noted in electrode performance due to variations in unfrozen water content in both thawed and frozen materials. In some situations the improvement in grounding conditions during thaw periods can be extended by use of conductive backfill. The lower freezing point of the backfill will also reduce electrode contact resistance caused by freezing around the metallic electrodes.

Over a period of time, salt very likely will move into the soil adjacent to the electrode backfill and therefore will increase the effective area of the ground electrode and in turn reduce the resistance values. The level of the backfill should be checked annually to insure adequate levels are maintained to replenish this loss due to seepage.

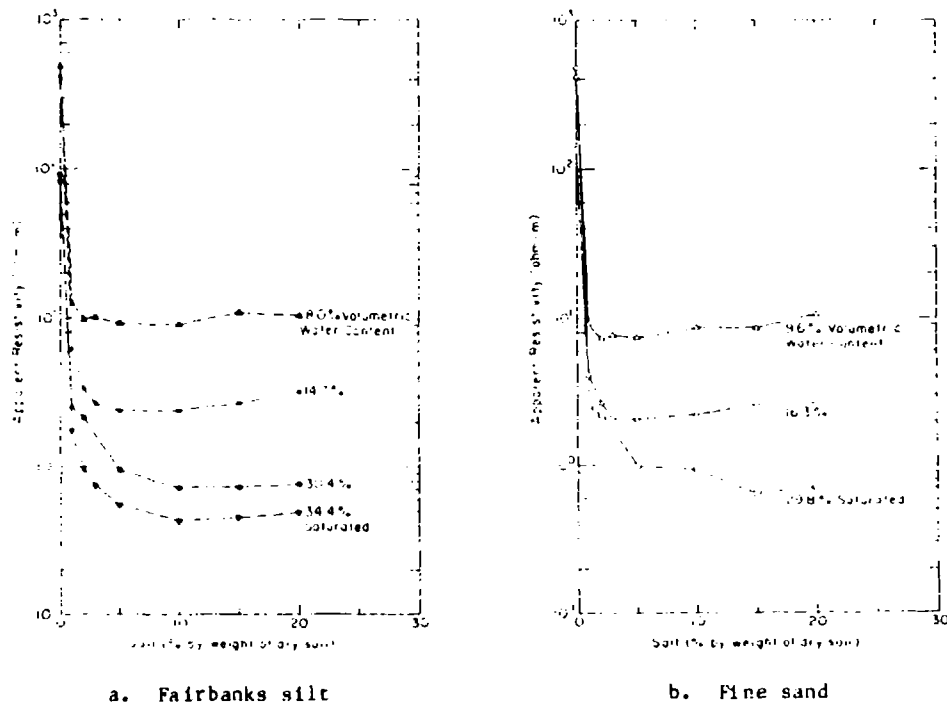


Figure 2-32. Apparent Resistivity for Two Soils at Various Moisture and Salt Contents

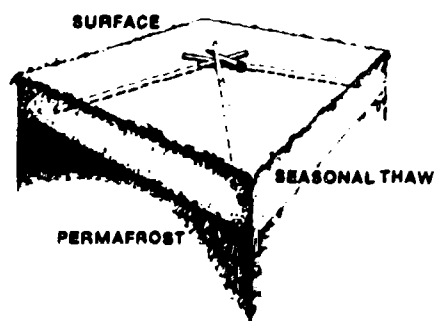


Figure 2-33. Configuration of Nearly Horizontal Electrodes Placed in the Thawed Active Layer

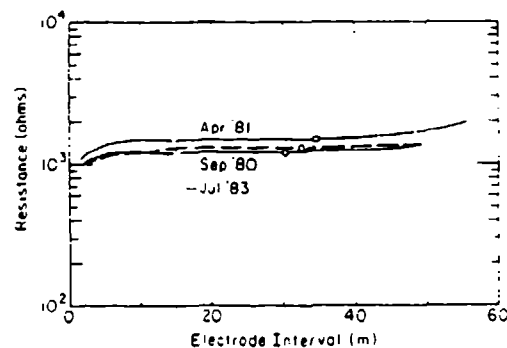


Figure 2-34. Resistance-to-Ground Curves for an Electrode Driven into Ice-Rich Silt

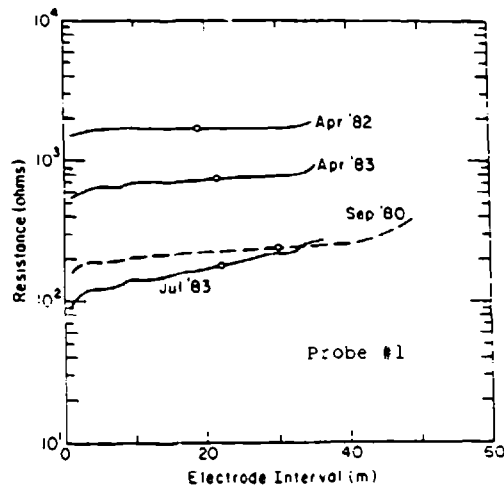


Figure 2-35. Resistance-to-Ground Curves for an Electrode Surrounded by a Backfill of Saturated Silt

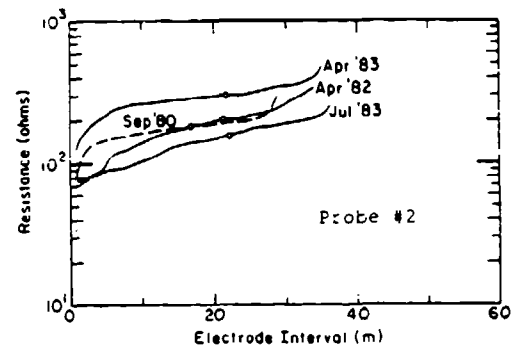


Figure 2-36. Resistance-to-Ground Curves for an Electrode Surrounded by a Water-Saturated Salt-Soil Backfill

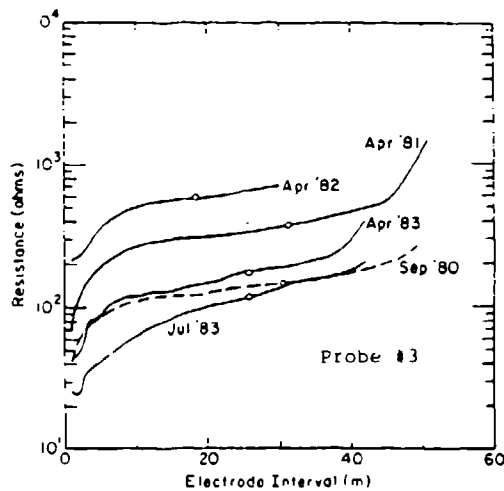


Figure 2-37. Resistance-to-Ground Curves for an Electrode Surrounded by a Water-Saturated Salt-Soil Backfill

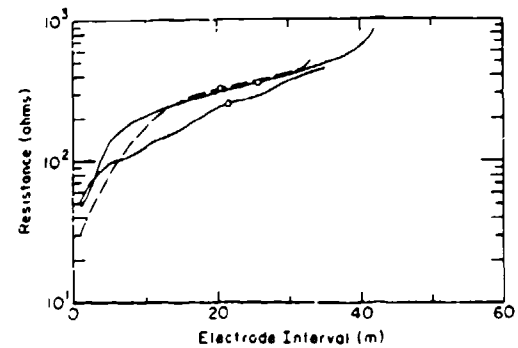


Figure 2-38. Resistance-to-Ground Curves for 3 Electrodes Placed in Holes Modified by Spring Charges and Filled with a Salt-Water Solution.



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## CHAPTER 3

## LIGHTNING PROTECTION SUBSYSTEM

3.1 THE PHENOMENON OF LIGHTNING.

Cumulonimbus clouds associated with thunderstorms are huge, turbulent air masses extending as high as 15 to 20 kilometers (9 to 12 miles) into the upper atmosphere. Through some means, not clearly understood, these air masses generate regions of intense static charge. These charged regions develop electric field gradients of hundreds, or perhaps thousands, of millions of volts between them. When the electric field strength exceeds the breakdown dielectric of air ( $\approx 3 \times 10^5$  volts/meter), a lightning flash occurs and the charged areas are neutralized.

Electric field measurements indicate that the typical thundercloud is charged in the manner illustrated by Figure 3-1 (3-1). A strong, negatively charged region exists in the lower part of the cloud with a counterbalancing positive charge region in the upper part of the cloud. In addition to these major charge centers, a smaller, positively charged region exists near the bottom of the cloud. Due to the strong negative charge concentration in the lower portion of the cloud, the cloud appears to be negatively charged with respect to earth -- except in the immediate vicinity underneath the smaller positive charge concentration.

Breakdown can occur between the charged regions within the cloud to produce intracloud lightning. It can also occur between the charged regions of separate clouds to produce cloud-to-cloud lightning. Intracloud and cloud-to-cloud discharges do not present a direct threat to personnel or structures on the ground and thus tend to be ignored in the design and implementation of lightning protection systems. However, calculations of the voltages which could be induced in cross-country cables by such discharges (3-2) indicate that they present a definite threat to signal and control equipments, particularly those employing solid state devices.

The cloud-to-ground flash is the one of primary interest to ground-based installations. By definition, such flashes take place between a charge center in the cloud and a point on the earth. This point on earth can be a flat plain, body of water, mountain peak, tree, flag pole, power line, residential dwelling, radar or communications tower, air traffic control tower, or multi-story skyscraper. In a given area, certain structures or objects are more likely to be struck by lightning than others; however, no object whether man-made or natural feature, should be assumed to be immune from lightning.

The high currents which flow during the charge equalization process of a lightning flash can melt conductors, ignite fires through the generation of sparks or the heating of metals, damage or destroy components or equipments through burning or voltage stressing, and produce voltages well in excess of the lethal limit for people and animals. The objective of all lightning protection subsystems is to direct these high currents away from susceptible elements or limit the voltage gradients developed by the high currents to safe levels.

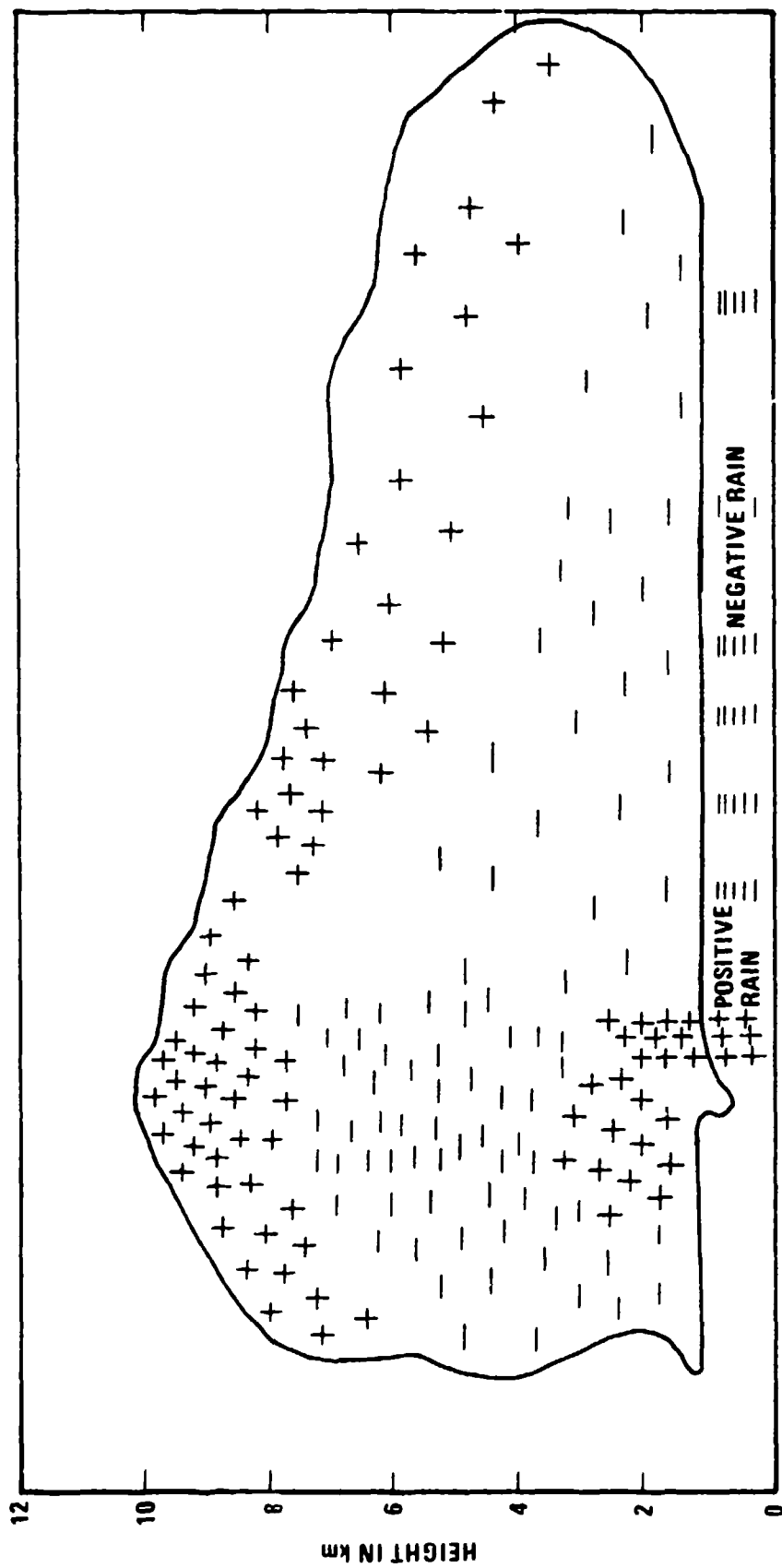


Figure 3-1. Charge Distribution in a Thundercloud

### 3.2 DEVELOPMENT OF A LIGHTNING FLASH.

As the charge builds up in a cloud, the electric field in the vicinity of the charge center builds up to the point where the air starts to ionize. A column of ionized air, called a pilot streamer, begins to extend toward earth at a velocity of about 160 kilometers per hour (100 miles per hour) (3-3). After the pilot streamer has moved perhaps about 30 to 45 meters (100 feet to 150 feet), a more intense discharge called a stepped leader takes place. This discharge lowers additional negative charge into the region around the pilot streamer and allows the pilot streamer to advance for another 30 to 45 meters (100 to 150 feet) after which the cycle repeats. The stepped leader progresses towards the earth in a series of steps with a time interval between steps on the order of 50 microseconds (3-4).

In a cloud-to-ground flash, the pilot streamer does not move in a direct line towards the earth but instead follows the path through the air that ionizes most readily. Although the general direction is toward the earth, the specific angle of departure from the tip of the previous streamer that the succeeding pilot streamer takes is rather unpredictable. Therefore, each 30 to 45 meter (100 to 150 foot) segment of the discharge will likely approach the earth at a different angle. This changing angle of approach gives the overall flash its characteristic zig-zag appearance.

Being a highly ionized column, the stepped leader is at essentially the same potential as the charged area from which it originates. Thus, as the stepped leader approaches the earth, the voltage gradient between the earth and the tip of the leader increases. The increasing voltage further encourages the air between the two to break down.

The final stepped leader bridges the gap between the downward progressing column and the earth or an extension of the earth such as a tree, building, or metal structure that is equipotential with the earth. While the stepped leader is approaching the earth, a positive charge equivalent to the negative charge in the cloud is accumulating in the general region underneath the approaching leader. Once the stepped leader contacts earth (or one of its extensions), the built-up positive charge in the earth flows rapidly upward through the ionized column established by the stepped leader to neutralize the strong negative charge of the cloud. This return current constitutes what is generally referred to as the lightning stroke. If additional pockets of charge exist in the cloud, these pockets may discharge through the ionized path established by the initial stroke. Continuous dart leaders proceed from a remaining charge pocket toward the earth down this path. Once the dart leader reaches the earth, another return stroke of positive charge propagates up the channel to neutralize the secondary charge in the cloud. This cycle may be repeated several times as succeeding charge centers in the cloud are neutralized.

### 3.3 INFLUENCE OF STRUCTURE HEIGHT.

Flashes to earth are normally initiated by a pilot streamer from the cloud. As the charged leader approaches the ground, the voltage gradient at the surface increases. Ultimately the voltage becomes high enough for an upward-moving leader to be induced. Over flat, open terrain, the length of the upward leader does not exceed a few meters before it unites with the downward leader to start the return stroke. However, structures or other extensions from the earth's surface experience intensified electric field concentrations at their tips. Consequently the upward leaders are generated while the downward leader is some distance away; the upward

leader can be several hundred meters long before the two meet. For very tall buildings, the upward leaders begin to form even before the downward leaders have begun to form within the cloud; such incidents are generally described as triggered lightning. Triggered lightning is not very common for structures less than 150 meters (500 feet) in height; as the height increases above this threshold, the proportion of triggered strikes increases rapidly (3-5).

### 3.4 STRIKE LIKELIHOOD.

The number of total flashes to which the structure is exposed is related principally to local thunderstorm activity. Local thunderstorm activity can be projected from isokeraunic maps similar to those shown in Figures 3-2 and 3-3. These maps show the number of thunderstorm days per year for various regions of the United States and the world. Additional maps of worldwide keraunic levels can be obtained from the World Meteorological Association (3-6).

A thunderstorm day is defined as a local calendar day on which thunder is heard irrespective of whether the lightning flashes are nearby or at some distance away. To an observer at a specific location, the average distance at which lightning may occur and thunder will be heard is about 10 km (6 miles) (3-5). Therefore, a thunderstorm day means that at least one lightning discharge has occurred within an area of about 300 square km (120 square miles) surrounding the position of the observer. The actual number of strikes in the immediate vicinity of the observer may be considerably higher or lower than the number of thunderstorm days might indicate, depending upon the duration and intensity of a specific storm or series of storms.

In spite of the relative inexactness of a prediction of a lightning strike to a specific object that is based on the keraunic level, the thunderstorm day is the only parameter related to lightning incidence that has been documented extensively over many years. Its primary value lies in the qualitative information which it provides. This information can be used to assist in the determination of whether lightning protection should be provided in those situations where there is serious doubt as to the relative need for such protection. For example, a particular facility may not be essential to the safety of aircraft, but the loss of the facility may cause traffic delay. In an area of frequent thunderstorms such as the west coast of Florida, for example, the number of outages in areas where there was no protection could be so high as to be unacceptable; in an area of few thunderstorms; e.g., Southern California or Alaska, the expected outage from lightning might be once every few years (which could be significantly less than outages for routine maintenance).

The number of lightning flashes per unit earth surface area increases with the number of thunderstorm days per year, though not linearly. Empirical evidence indicates that the number of flashes per square kilometer,  $\sigma_y$ , can be reasonably predicted from (3-5):

$$\sigma_y = 0.007 T_y^2 \quad (3-1)$$

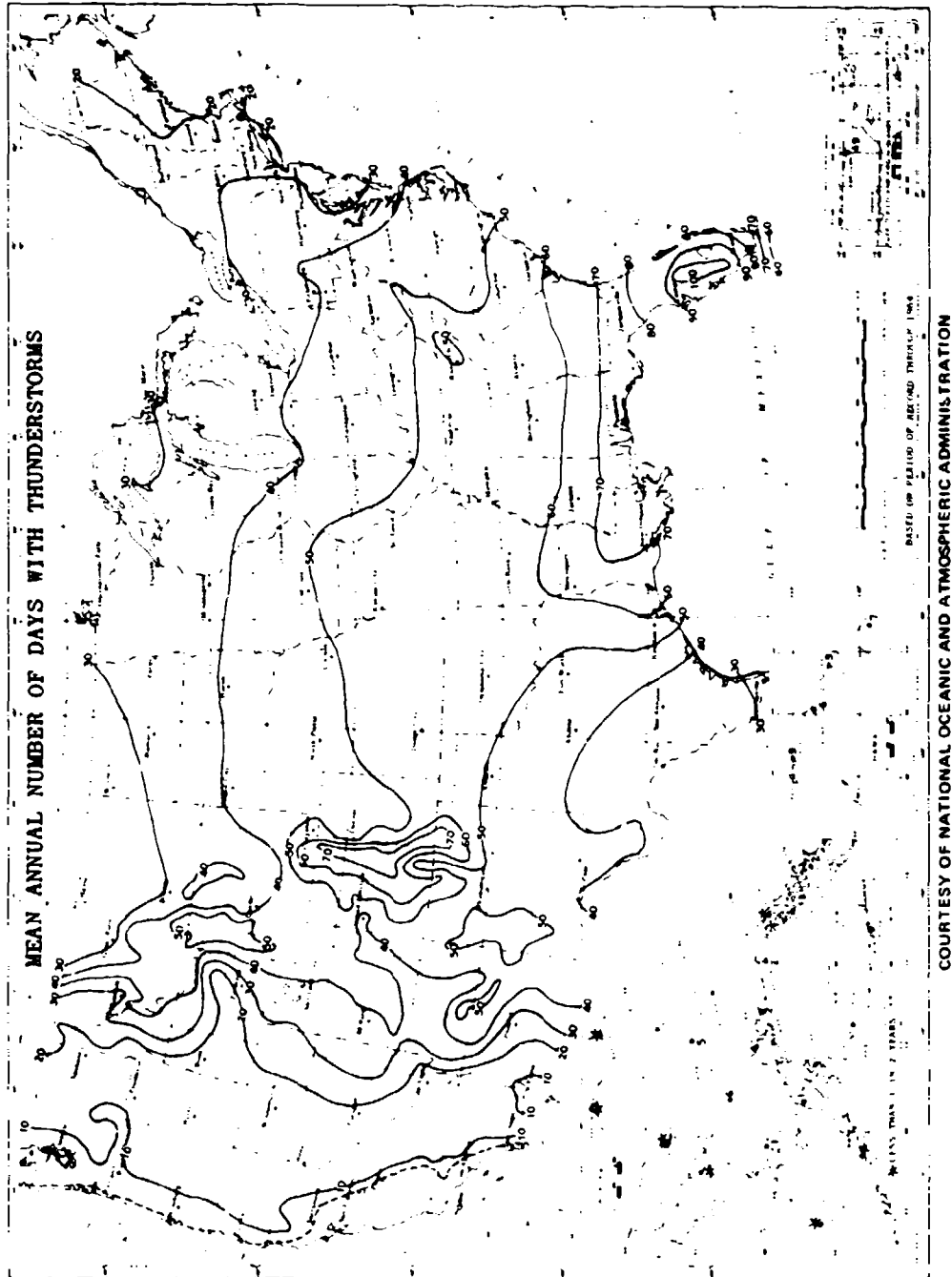


Figure 3-2. Mean Number of Thunderstorm Days Per Year for the United States

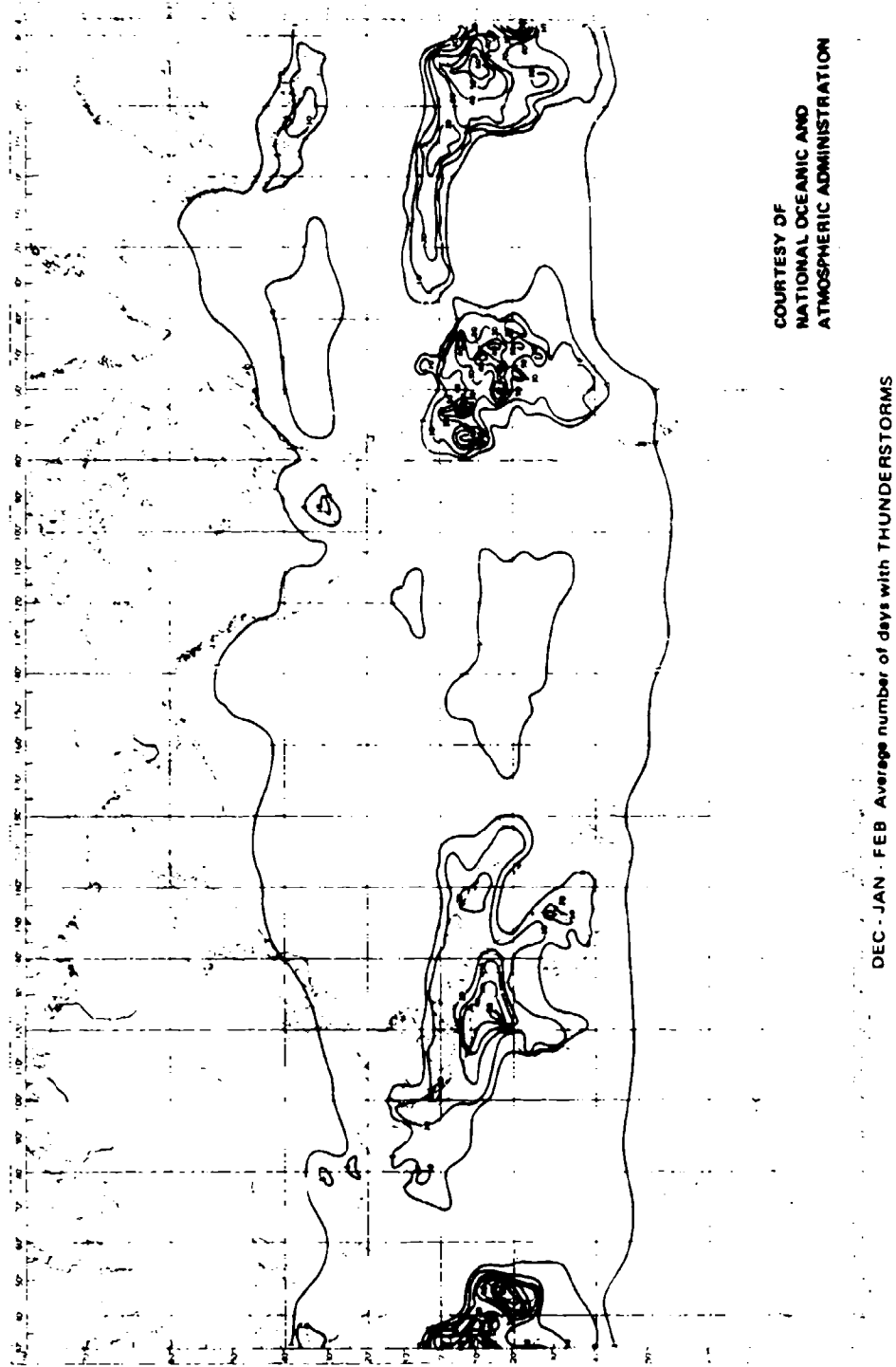


Figure 3-3. Worldwide Isokeraunic Map (Sheet 1 of 4)

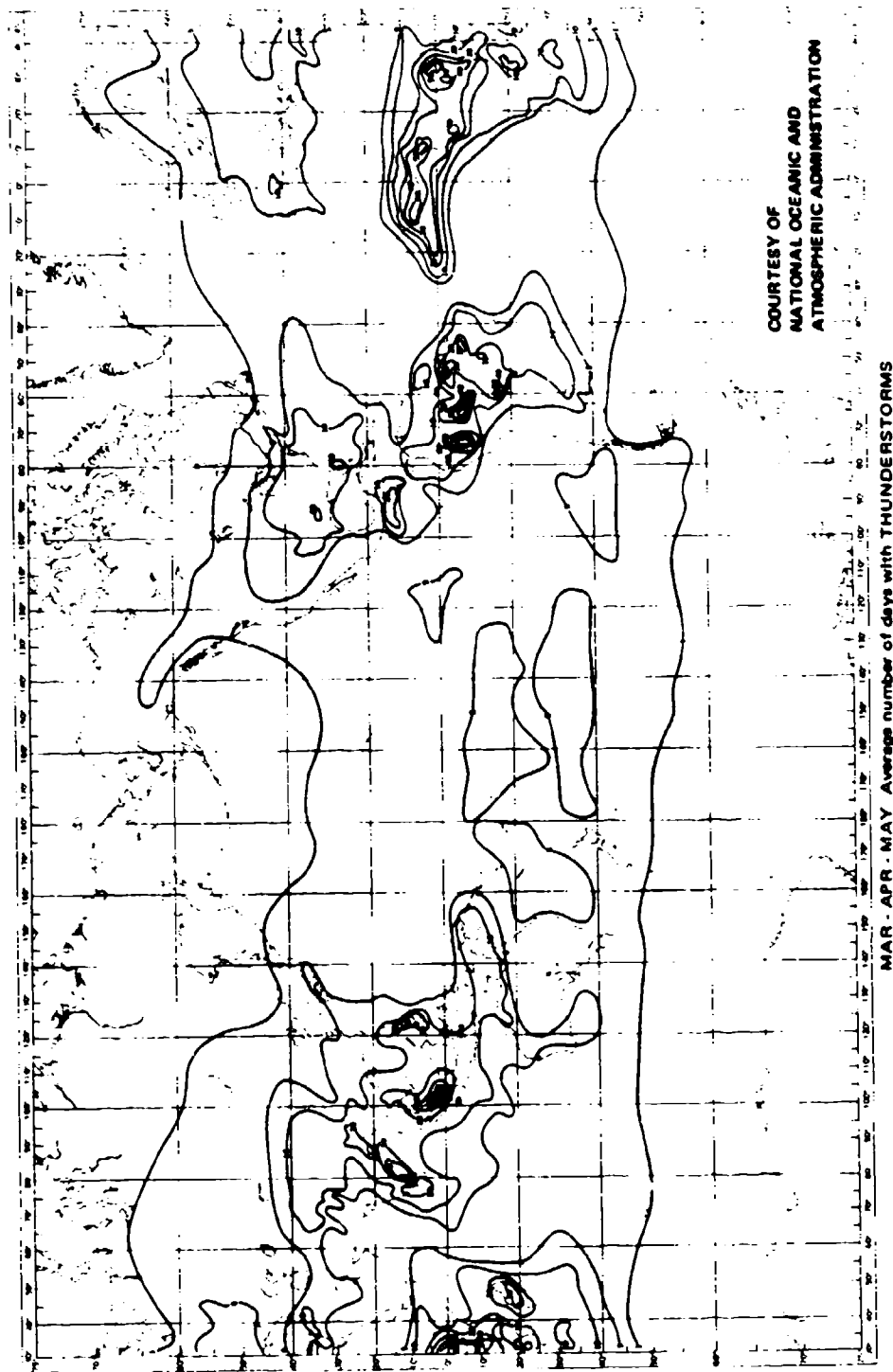


Figure 3-3. Worldwide Isokeraunic Map (Sheet 2 of 4)



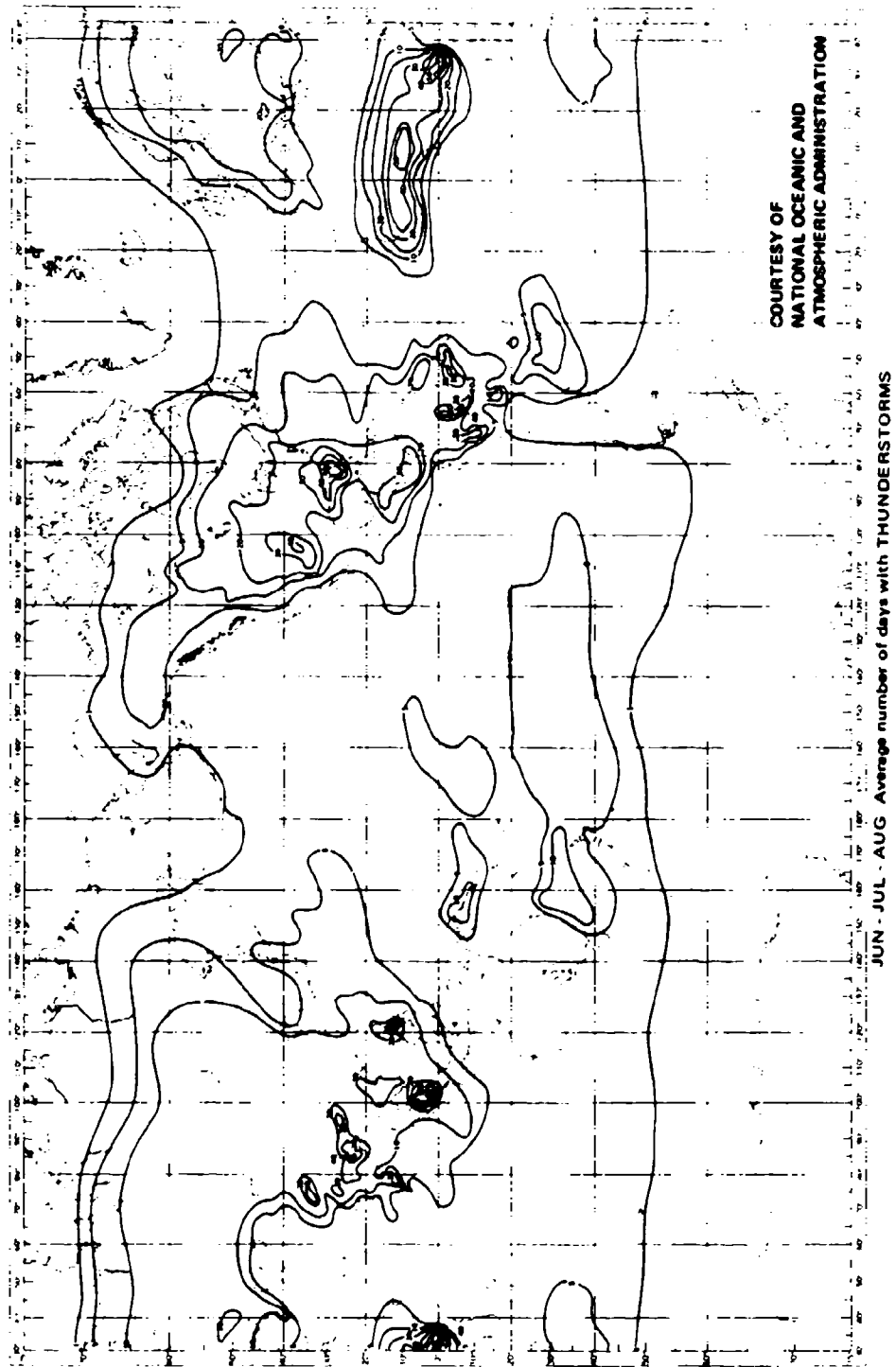


Figure 3-3. Worldwide Isokeraunic Map (Sheet 3 of 4)

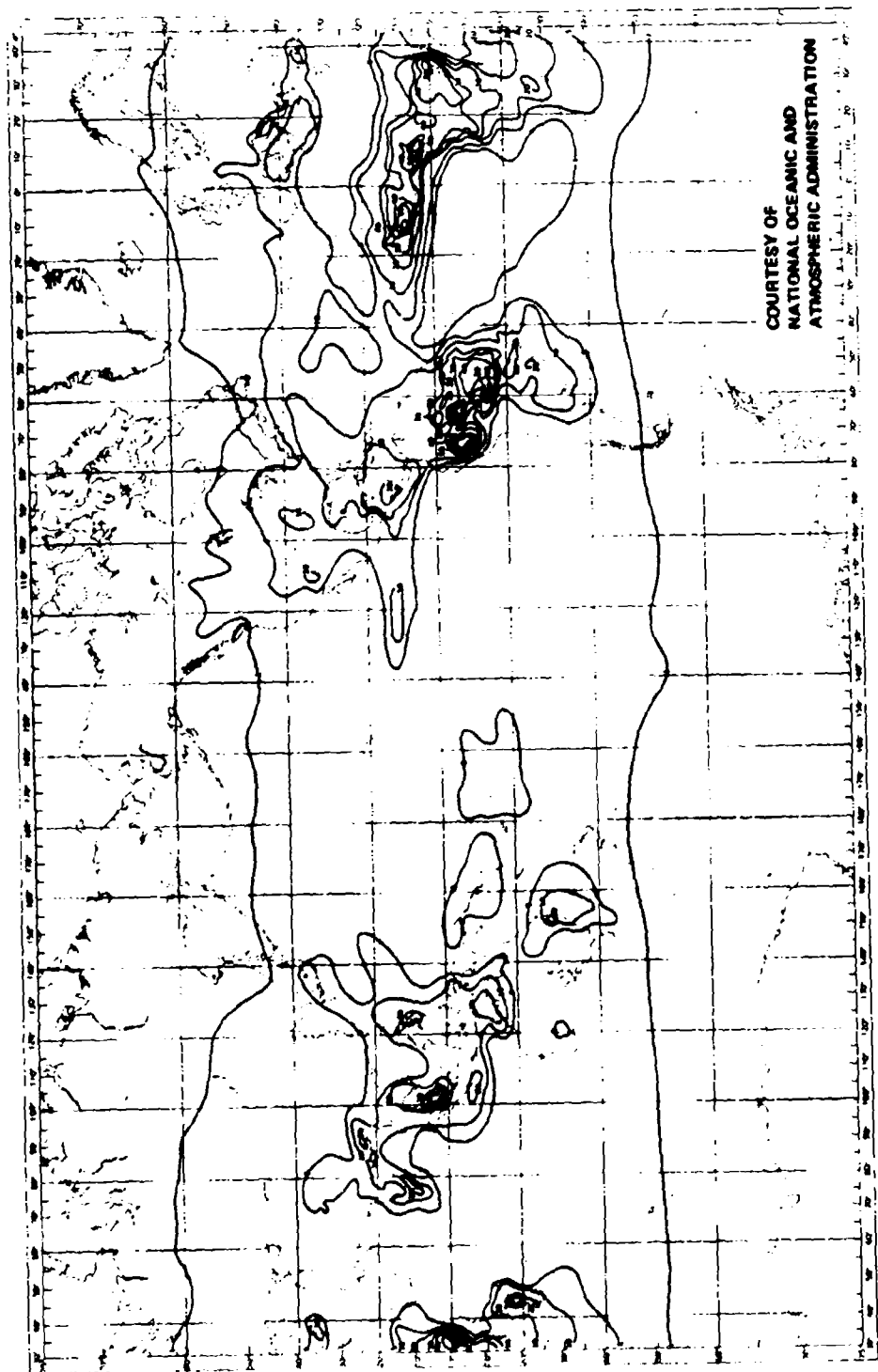


Figure 3-3. Worldwide Isokeraunic Map (Sheet 4 of 4)

where  $T_y$  is the number of thunderstorm days per year. Out of the total number of flashes per unit area, the number of discharges increases with increasing geographical latitude (3-7). The proportion,  $p$ , of discharges that go to ground in relation to the geographical latitude,  $\Lambda$ , can be represented (3-8) as:

$$p = 0.1 (1 + (\Lambda/30)^2) \quad (3-2)$$

Thus in a given location the flash density,  $\sigma_{yg}$ , i.e., the number of discharges to earth per square kilometer per year, is:

$$\sigma_{yg} = p\sigma_y \quad (3-3)$$

To calculate  $\sigma_{yg}$ , for a specific location, first determine  $T_y$  from the isokeraunic map of Figure 3-2 to Figure 3-3. For estimation purposes, the number of thunderstorm days at points between lines may be determined by interpolation. Using this value of  $T_y$ , calculate the total flash density with Equation 3-1. Next obtain the geographical latitude of the site from a map of the area and calculate  $p$  from Equation 3-2. Then determine the number of strikes to earth per year per square kilometer with Equation 3-3.

**3.5 ATTRACTIVE AREA.** The concept of attractive area reflects the principle that an object extending above its surroundings is more likely to be struck by lightning than its actual cross-sectional area might otherwise indicate. For example, thin metallic structures such as flag poles, lighting towers, antennas, and overhead wires offer a very small cross-sectional area relative to the surrounding terrain but ample evidence exists to show that such objects apparently attract lightning.

#### 3.5.1 Structures Less Than 100 Meters High.

For structures less than 100 meters (330 feet) in height, and which therefore do not normally trigger lightning, the number of strikes increases according to a power of  $h$ , the structure height. An expression that represents the attractive radius,  $r_a$ , in meters of a structure is (3-5);

$$r_a = 80 \sqrt{h} (e^{-0.02h} - e^{-0.05h}) + 400 (1 - e^{-0.0001h^2}) \quad (3-4)$$

where  $h$  is in meters. For a structure 10 meters high, Equation 3-4 gives an attractive radius of 57.7 meters; similarly, the attractive radius for a 100-meter high structure is 358 meters. The attractive area,  $A_a$ , is  $\pi r_a^2$ . Thus  $A_a$  for a 10-meter structure is approximately 0.01 square kilometer, while the attractive area of a 100-meter structure is 0.4 square kilometer.

Equation 3-4 has been found to adequately describe the number of strikes to objects which are not tall enough to trigger lightning. For taller structures, a multiplication factor (3-5)

$$F_T = 1 + 2 (9-1500/h) \quad (h \text{ in meters}) \quad (3-5)$$

should be applied to Equation 3-4. The experimental data to justify the use of Equation 3-5 for structures greater than 400 meters (1300 feet) is sketchy. However, since structures even approaching this height are not expected to be of primary concern, Equations 3-4 and 3-5 are expected to be adequate for most design purposes.

Large flat buildings that do not extend above the median treetop level in the general area will have an attractive area that is essentially the area of the roof (assuming the roof covers the entire structure). If the building is several stories high such that it appreciably extends above the prevailing terrain, then its attractive area is its roof area plus that portion of the attractive area not already encompassed by the roof. Figure 3-4 illustrates the method for calculating the attractive area of a rectangular structure of length,  $l$ , and width,  $w$ . The roof area is given by  $l \times w$ . The additional attractive area resulting from the height of the building is readily determined by recognizing that the areas contributed by the four corners of the building equal a circle of radius,  $r_a$ . Both ends of the structure (dimension  $w$ ) contribute the area of  $2 \pi r_a^2$ ; the sides contribute  $2l \pi r_a$ . The total attractive area is the sum of the roof area ( $lw$ ), the corners ( $\pi r_a^2$ ), the ends ( $2 \pi r_a^2$ ), and the sides ( $2l \pi r_a$ ) to produce a total of

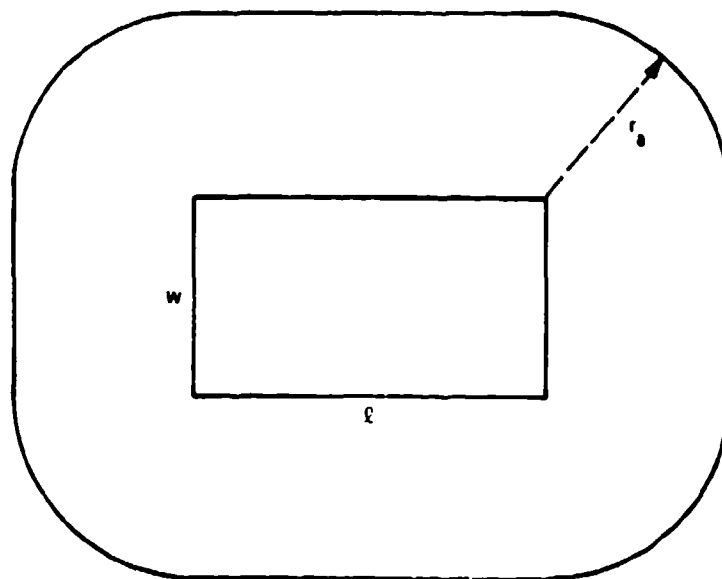
$$A_a = wl + \pi r_a^2 + 2r_a (\pi + l). \quad (3-6)$$

Figure 3-5 indicates that the height to be used in calculating the attractive area of a tall structure should be the height that the structure extends above the effective (i.e., the level that earth charges would rise to if the building were not there) levels of the earth. On open, level terrain the height,  $h$ , would be the full height of the roof from grade level.

The number of flashes which can be expected to strike a given structure is equal to the product of the flash density,  $\sigma_{yg}$ , times the attractive area,  $A_a$ , of the structure. For example, suppose the relative likelihood of a lightning strike to a low, flat structure 100 meters on a side, located in Nashville, TN, is desired. From Figure 3-2,  $T_y$  is determined to be approximately 54 thunderstorm days per year. The flash density as given by Equation 3-1 is 20.4 flashes/km<sup>2</sup>/year. The proportion of those flashes that are discharges to earth is 24.4 percent (from Equation 3-2) since the latitude is 36 degrees. Thus approximately 5 flashes/km<sup>2</sup>/year to earth can be expected. Within the area of the structure (0.01 km<sup>2</sup>) there will be only 0.05 strikes per year on the average, or there is a 1 in 20 chance of being struck by lightning in a given year. For the same structure in Southern California, only a 1 in 330 likelihood of a strike would be expected in a given year.

### 3.5.2 Cone of Protection.

This ability of tall structures or objects to attract lightning to themselves serves to protect shorter objects and structures. In effect, a taller object establishes a protected zone around it. With this protected zone, other shorter structures and objects are protected against direct lightning strikes. As the heights of these shorter objects increase, the degree of protection decreases. Likewise, as the separation between tall and short structures increases, the protection afforded by the tall structure decreases. The protected space surrounding a lightning conductor is called the zone (or cone) of protection.



TOTAL ATTRACTIVE AREA:  $A_s = wl + \pi r_s^2 + 2r_s(w + l)$

$r_s$  IS DETERMINED BY EFFECTIVE HEIGHT,  $h$ . SEE EQUATIONS 3-4 AND 3-5.

Figure 3-4. Attractive Area of a Rectangular Structure

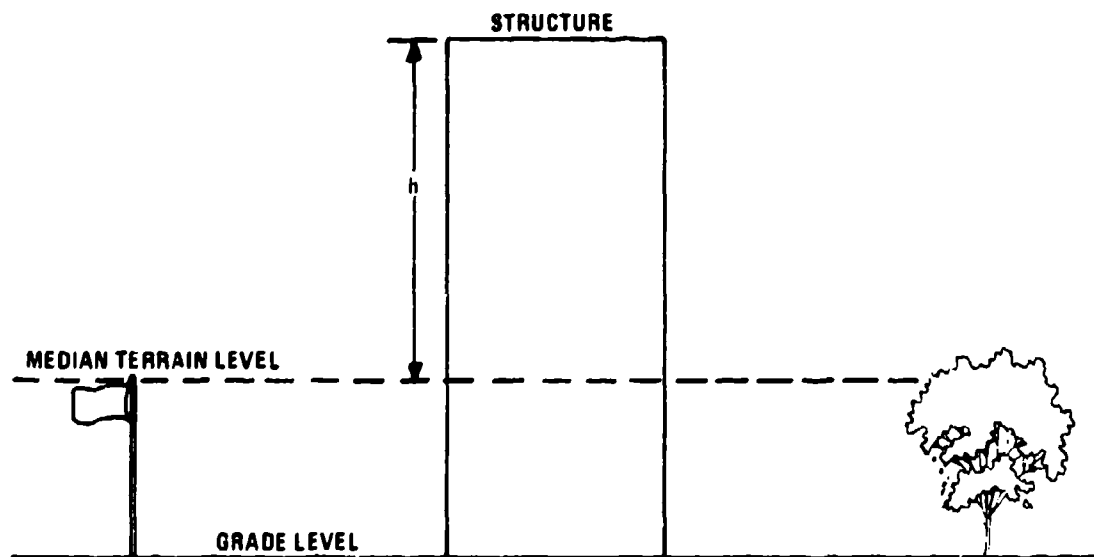


Figure 3-5. Effective Height of a Structure

The zone of protection provided by a grounded vertical rod or mast is conventionally defined as the space enclosed by a right circular cone with its axis coincident with the mast and its apex at the top of the mast as illustrated by Figure 3-6(a). Similarly, the zone protected by a grounded horizontal overhead wire is defined as a triangular prism with its upper edge along the wire as illustrated in Figure 3-6(b). In either case, the zone (or cone) of protection is expressed as the ratio of the horizontal protected distance,  $D$ , to height,  $H$ , of the mast or wire. This ratio is the tangent of the shielding angle,  $\alpha^*$ . Some commonly recommended zones of protection and the associated shielding angles are illustrated in Figure 3-7.

The NFPA Lightning Protection Code (3-9) recommends that a 1:1 zone of protection ( $\alpha = 45^\circ$ ) be provided in important areas while a 2:1 zone ( $\alpha = 63^\circ$ ) is acceptable for less important areas. The British Standard Code of Practice (3-10) states that a shielding angle of 45 degrees provides an acceptable degree of protection for ordinary structures, but that for structures with explosive or high flammable contents the shielding angle should not exceed 30 degrees.

Although the existence of a 1:1 zone of protection does not absolutely guarantee immunity to lightning, documented cases of the 1:1 zone being violated are very few. Thus for all facilities except those associated with the storage of explosives or fuels, a 1:1 zone of protection can safely be used as a basis of design of lightning protection systems. As such, C-E facilities or equipments (antennas, etc.) located entirely within the 1:1 zone of protection generally are not required to have separate air terminals. This does not eliminate the need to ground metal shelters or to meet the grounding requirements of the subsystems which comprise the facility ground system. If more than one rod or wire is used, the protected zone is somewhat greater than the total of all of the 1:1 zones of the rods or wires considered individually. For adjacent structures, the Codes specify that a 2:1 zone of protection may be assumed for the region between the structures.

Large structures with flat or gently sloping roofs do not lend themselves to the straightforward application of the 1:1 or 2:1 zone of protection principles. To establish even 2:1 type coverage on large buildings, exceptionally tall air terminals would be required. Experience, however, shows that extremely tall terminals are not needed for effective protection. Both the NFPA Lightning Protection Code and UL Master Labeled Protection System (3-11) specify air terminals that extend from 10 to 36 inches above the object to be protected. (The British Standard Code of Practice does not require the use of air terminals at all.)

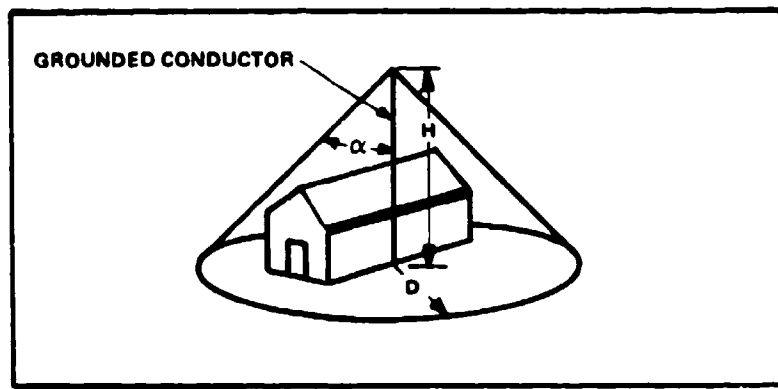
### 3.6 LIGHTNING EFFECTS.

#### 3.6.1 Flash Parameters.

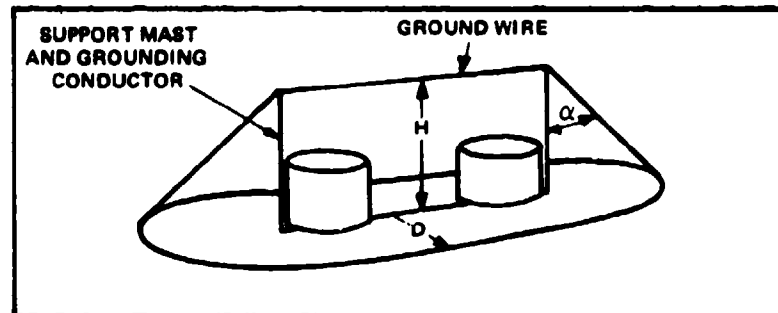
During the short interval of a lightning flash, several discharges occur. The sequence of events in a multiple-stroke flash is illustrated in Figure 3-8. The initial path for the discharge is established in 50 microseconds. Intermediate return stroke currents of about 1 kA follow the initial return stroke and last for a few milliseconds. Subsequent strokes occur at intervals of 50 to 60 milliseconds. The return stroke interval may include a continuing current of 100 A or so which flows for several milliseconds or until the start of the next return stroke.

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\*The shielding angle is defined as the angle between the surface of the cone and a vertical line through the apex of the cone, or between the side of the prism and the vertical plane containing the horizontal wire.



(a) CONE OF PROTECTION PROVIDED BY A VERTICAL GROUNDED CONDUCTOR OF HEIGHT H.



(b) ZONE OF PROTECTION PROVIDED BY A HORIZONTAL AERIAL GROUND WIRE AT HEIGHT H.

Figure 3-6. Zones of Protection Established by a Vertical Mast and a Horizontal Wire

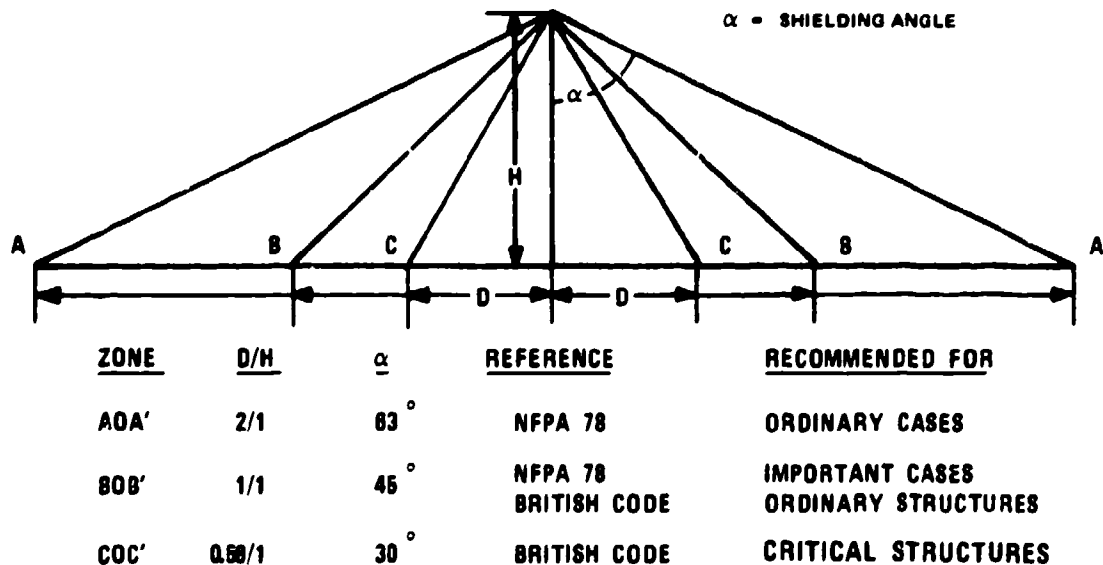


Figure 3-7. Some Commonly Used Lightning Shielding Angles

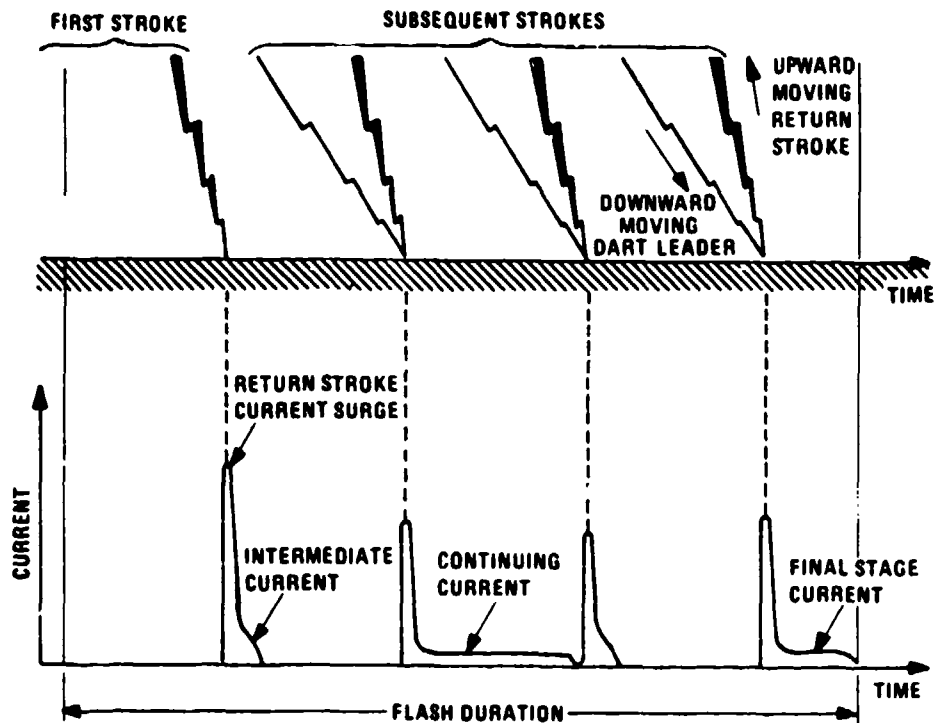


Figure 3.8. Illustration of Process and Currents Which Occur During a Lightning Flash to Ground

The lightning discharge involves the transfer of large amounts of electric charge between the cloud and the earth. The typical flash transfers 15 to 20 coulombs (C) (1 coulomb equals  $6.2 \times 10^{18}$  electrons) with some flashes involving as much as 400 coulombs of charge. The energy per flash of lightning has been estimated to be as high as  $10^8$  watt-seconds. Table 3-1 summarizes the range of values for selected lightning parameters.

### 3.6.2 Mechanical and Thermal Effects.

The fast rise time, high peak amplitude current of the stroke can produce severe mechanical, thermal, and electrical effects. The damage caused by these currents to objects in the discharge path is closely related to the relative conducting power of the object. For example, metals generally receive a discharge with little damage. In most cases, even slender conductors such as telephone and electric power cables handle the current without fusing (melting) except at the point where the current enters or leaves the metal (where severe damage may occur). Very strong discharges of high peak current ( $> 40$  kA) and high coulomb values ( $> 200$  C), however, can melt or burn holes in solid metal plates. This burning effect is not usually of primary concern for a typical building or structure because, if an adequate protection system is installed, the principle effect will be a small deformation at the tip of a lightning rod or a small melted area on the intercepting cable. Such effects are of more concern where flashes to airplanes occur because such burning can perforate the fuselage to cause loss of pressurization or penetrate the skin of fuel tanks and possibly ignite fuels. The burning or melting also presents a threat to exposed tanks of volatile gases or fuels on the ground.



Table 3-1

## Range of Values for Lightning Parameters (3-5)

<u>Parameter</u>	<u>Minimum</u>	<u>Typical</u>	<u>Maximum</u>
Number of return strokes per flash	1	2 to 4	26
Duration of flash(s)	0.03	0.2	2
Time between strokes (ms)	3	40 to 60	100
Peak current per return stroke (kA)	1	10 to 20	250
Charge per flash (C)	1	15 to 20	400
Time to peak current ( $\mu$ s)	<0.5	1.5 to 2	30
Rate of rise (kA/ $\mu$ s)	<1	20	210
Time to half-value ( $\mu$ s)	10	40 to 50	250
Duration of continuing current (ms)	50	150	500
Peak continuing current (A)	30	150	1600
Charge in continuing current (C)	3	25	330

Because of the duration of the currents that flow for the extended intervals between return strokes, they are most likely to cause damage by melting or igniting solid materials. In contrast, the short-duration high-current peaks tend to tear or bend metal parts by the electromagnetic forces that develop in proportion to the square of the instantaneous current. Though potentially hazardous, the damage caused by mechanical forces in metallic conductors is generally of secondary importance in most situations. However, because of the presence of these mechanical forces, it is necessary that lightning rods, down conductors, and other elements of the protection system be securely anchored.

On the other hand, when insulating or semi-insulating material receives a discharge, an explosive reaction may occur with severe damage. Trees, for instance, whether dry or green, are in many cases split or stripped of their bark, and the damage can extend underground to their roots. Related damage may occur to other unprotected wooden structures or objects such as flag poles, masts, or light supports, and electric and telephone poles. When lightning strikes a wooden building, the stroke seeks out the lowest impedance path to earth which is probably through the electric wiring or water pipes. Often in order to reach these metallic paths, the discharge must pass through some type of wooden barrier. In penetrating such barriers, extensive explosive damage usually results.

Brick, concrete, marble, and other masonry materials are also frequently shattered or broken loose at the point where the discharge passes through them. Such damage will occur where structural steel support members or steel reinforcing rods are encased in concrete or sheathed in brick or marble and the structure has an inadequate protective system. The explosive effect can dislodge materials with considerable force -- force sufficient to hurl relatively large pieces several meters. One explanation of the explosive force is that it is the result of the virtually instantaneous vaporization of the water present in the wood or entrapped in the masonry materials.

### 3.6.3 Electrical Effects.

Lightning discharges to or near the buildings and structures frequently cause damage to electrical and electronic equipment. Melting or burning of conductors occurs at the point of interception of the stroke. The voltages developed by the fast risetime, high amplitude current pulse are frequently high enough to break down insulation, pose personnel hazards, and cause component and device failure. These voltages are produced by:

- a.  $IZ$  (current  $\times$  impedance) drop resulting from the lightning pulse traveling down power lines or signal lines, through structural members, along down conductors or overhead ground wires or through the resistance of the earth connection;
- b. Magnetic induction; and
- c. Capacitive coupling.

Lightning surges in power, signal, and control circuits are generally the result of some combination of these three components.

#### 3.6.3.1 Conductor Impedance Effects.

Because of the fast risetime (1 to 2  $\mu\text{sec}$ ) and high amplitude (10 to 20 kA) characteristics of the current pulse produced by the lightning discharge, the inductance and resistance of even relatively short conductors causes extremely high voltages to be developed on the conductor. The voltages frequently are high enough to exceed the breakdown potential of air or other insulation materials and cause flashover to other conductors or breakdown of insulation. The resistive IR drop generated by 20 kA in a 30 meter (100 feet) run of down conductor conforming to NFPA-78 ( $2.88 \times 10^{-4} \Omega/\text{m}$ ) will be

$$V = 2 \times 10^4 \times 2.88 \times 10^{-4} \times 30 = 173 \text{ volts} \quad (3-7)$$

which is not sufficient to cause flashover or to pose a serious threat to personnel.

For a down conductor length of 30 meters (100 feet), the smallest copper conductor meeting the minimum requirements of the Lightning Protection Code or the UL Master Labeled Lightning Protection System has a diameter of 0.894 cm (0.352 inches). Assuming that the conductor is a straight round wire, the inductance can be determined from (see Section 5.2.2.3):

$$L = 0.002 \ell \left( 2.303 \log \frac{4\ell}{d} - 0.75 \right), \quad (3-8)$$

where  $L$  is the total inductance in microhenries,  $\ell$  is the length in cm, and  $d$  is the diameter in cm. A 30-meter length of conductor will exhibit an inductance of 52.5 microhenries.

The voltage,  $V$ , developed across an inductance is given by

$$V = L \, di/dt, \quad (3-9)$$

where  $L$  is the inductance in henries and  $di/dt$  is the rate of change of the current through the inductor in amperes per second. From Table 3-1, the rate of rise of the typical lightning stroke is 20 kA/ $\mu$ s which corresponds to a  $di/dt$  of  $2 \times 10^{10}$  amps/second. Thus the voltage developed by the discharge pulse through the 30-meter (100 foot) downconductor is

$$V = 5.25 \times 10^{-5} \times 2 \times 10^{10} = 1.05 \times 10^6 \text{ volts.} \quad (3-10)$$

Although the duration of this voltage is typically less than 2 microseconds, the voltage generated is high enough to cause flashover to conducting objects located as much as 35 cm (14 in.) away from the down conductor. It is for this reason that metallic objects within 6 feet of lightning down conductors should be electrically bonded to the down conductors.

### 3.6.3.2 Induced Voltage Effects.

In addition to the lightning effects discussed above, circuits not in direct contact with the lightning discharge path can experience damages even in the absence of overt coupling by flashover. Because the high current associated with a discharge exhibits a high rate of change, voltages are electromagnetically induced on nearby conductors. Experimental and analytical evidence (3-12) shows that the surges thus induced can easily exceed the tolerance level of many components, particularly solid state devices. Surges can be induced by lightning current flowing in a down conductor or structural member, by a stroke to earth in the vicinity of buried cables, or by cloud-to-cloud discharges occurring parallel to long cable runs, either above ground or buried (3-2).

Consider a single-turn loop parallel to a lightning down conductor such as that shown in Figure 3-9. The voltage  $E$  magnetically induced in the loop is related to the rate of change of flux produced by the changing current in the down conductor (see Section 6.2.2.1). The voltage induced in the loop is dependent upon the dimensions of the loop ( $\ell$ ,  $r_2 - r_1$ ), its distance from the down conductor ( $r_1$ ), and the time rate of change of the discharge current ( $di/dt$ ). Figure 3-10 is a plot of normalized voltage per unit length that would be developed in a single turn loop of various widths.

These results suggest the steps that should be taken to minimize the voltage induced in signal, control, and power lines by lightning discharges through down conductors. First, since no control can be exercised over  $di/dt$  because it is determined by the discharge itself,  $E$  must be reduced by controlling  $\ell$ ,  $r_1$  and  $r_2$ . The variable  $\ell$ , is a measure of the distance that the loop runs parallel to the discharge path; thus, by restricting  $\ell$ , the induced  $E$  can be minimized. Thus cables terminating in devices or equipments potentially susceptible to voltage surges should not be run parallel to conductors carrying lightning discharge currents if at all possible. If parallel runs are unavoidable, Figure 3-10 also shows that the distance,  $r_1$ , between the loop and the lightning current path should be made as large as possible.

Another observation to be made from Figure 3-10 is that  $r_2$  minus  $r_1$  should be as close as possible to zero. In other words, the distance between the conductors of the pickup loop should be minimized. One common way of reducing this distance is to twist the two conductors together such that the average distance from each conductor to the discharge conductor is the same.

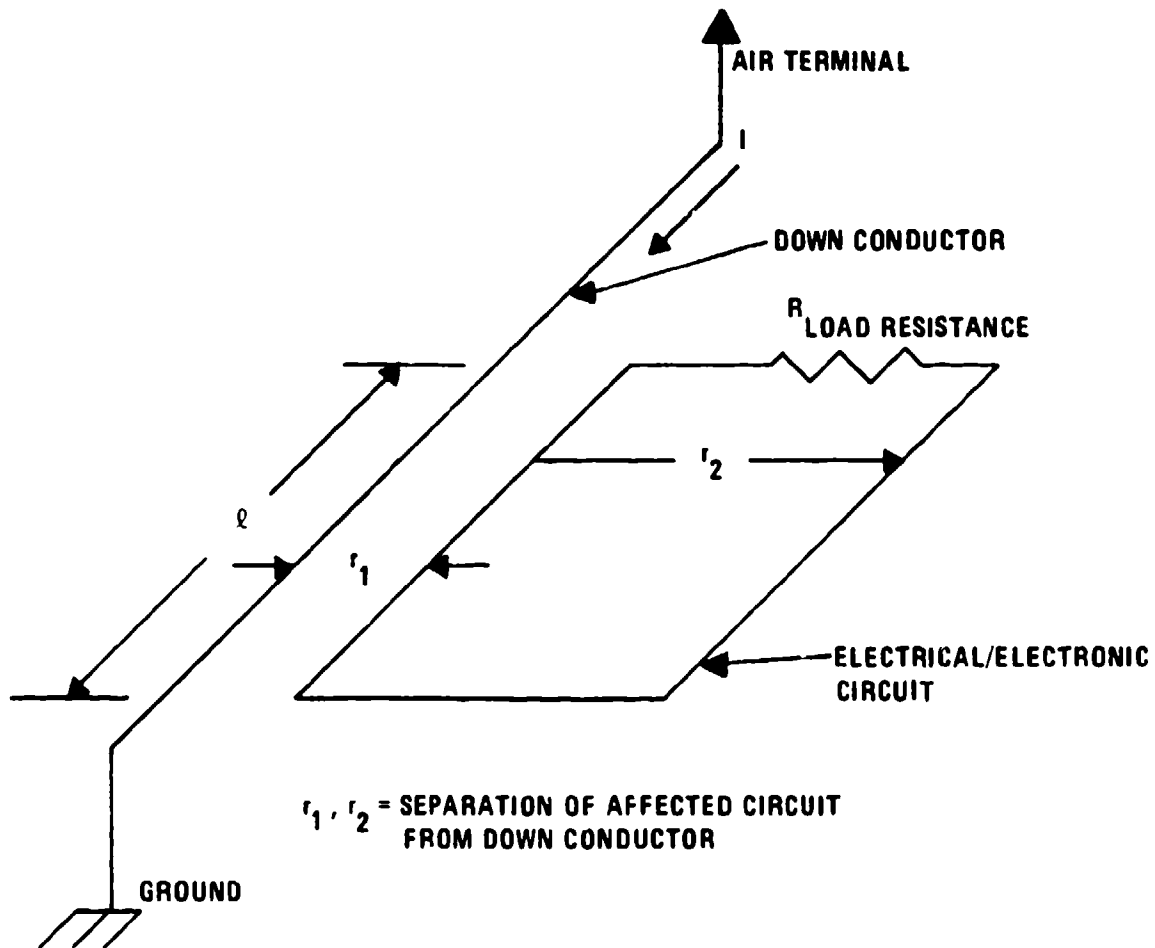


Figure 3-9. Inductive Coupling of Lightning Energy to Nearby Circuits

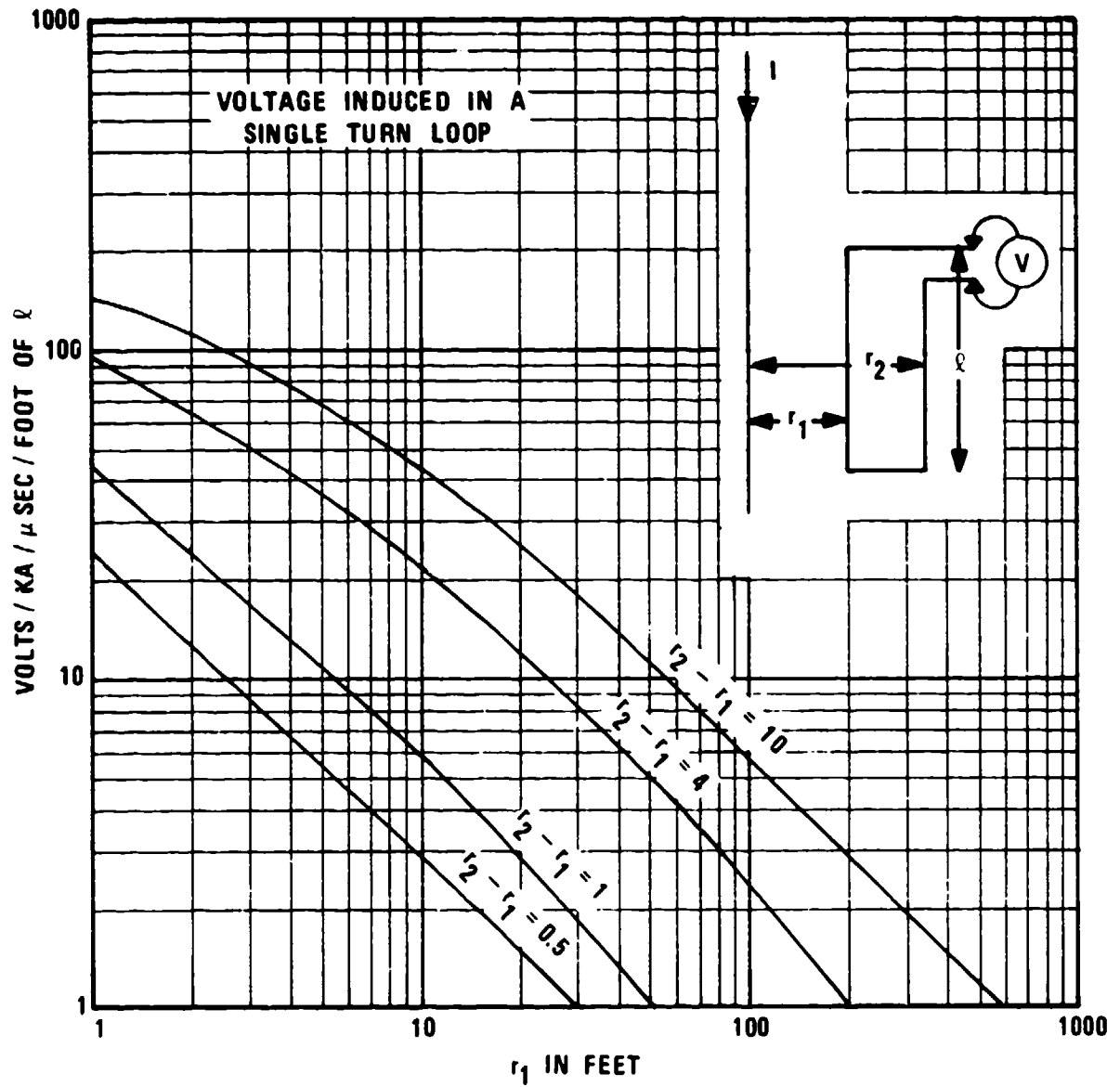


Figure 3-10. Normalized Voltage Induced in a Single-Turn Loop by Lightning Currents

Another protective measure is to reduce the flux density within the pickup loop by providing magnetic shielding. Because the coupling field is primarily magnetic in nature, a shielding material having a high permeability such as iron or nickel should be used. For shielding against lightning-produced fields, steel conduit or cast iron pipe are much more effective than aluminum or other non-ferrous materials.

### 3.6.3.3 Capacitively-Coupled Voltage.

Prior to the lightning discharge, an electric charge slowly accumulates on earth-based objects in the vicinity of the electrified clouds. This increase in charge occurs slowly enough so that the potential of grounded conductors does not change appreciably with respect to the earth, even when the impedance to ground is high. When the lightning stroke terminates on a structure or other point having contact with the earth as illustrated in Figure 3-11, the charge on all grounded objects nearby suddenly becomes redistributed. The redistribution of charge produces a current flow through the grounding impedance of the grounded objects and produces a voltage across that impedance.

Referring to Figure 3-11, the voltage between the conducting objects and the ground can be expressed as

$$E = \frac{Q}{C} e^{-t/RC} \quad (3-11)$$

where  $Q$  is the stored charge in coulombs,  $C$  is the total capacitance to ground in farads,  $R$  is the effective resistance to ground in ohms, and  $t$  is the elapsed time in seconds from the occurrence of the stroke.

Equation 3-11 shows that if the product  $RC$  is small, the exponential term will be large (for a time  $t$  on the order of 10  $\mu$ s), thus making the voltage capacitively induced on any reasonably well-grounded object quite small for a typical lightning stroke.

### 3.6.3.4 Earth Resistance.

Consider a facility such as the one illustrated in Figure 3-12, that has more than one possible electrical path to earth. For example, a ground rod is driven into the earth at the transformer pole or at the service entrance to Building 1. The resistance,  $R_{G1}$ , of this rod could be 25 ohms or higher and still conform to NEC requirements. Metal utility pipes such as water lines generally offer a relatively low resistance (labeled  $R_{G2}$ ) to earth. (In soils of high resistivity the point of effective contact between utility pipes may be an appreciable distance from the facility.) Empirical data indicates that the grounding resistance offered by water pipes is on the order of 1 to 3 ohms. If the electrical ground is not connected to the water pipe, a lightning strike to the ground wire of the electrical distribution system could produce a potential difference high enough to possibly produce an arc between the electrical ground (including the equipment cabinet and the building's structure, if connected) and the utility piping. A definite personnel hazard would then exist because of the high voltage that would be developed between the equipment and building ground and pipes. Because of this reason as well as the requirement to prevent analogous hazards from existing during power system faults, MIL-STD-188-124A requires electrical safety grounds be connected to the metallic water system in the building and recommends they also be directly connected to the ground rod at the transformer.

If  $R_{G1}$  is 25 ohms while  $R_{G2}$  is only 1 ohm or so, then a lightning strike as indicated could easily cause the potential of the overhead ground wire to become high enough to produce an arc across the transformer windings and insulators. Since the low voltage secondary side offers a lower impedance to earth, it is the preferred path for the discharge.

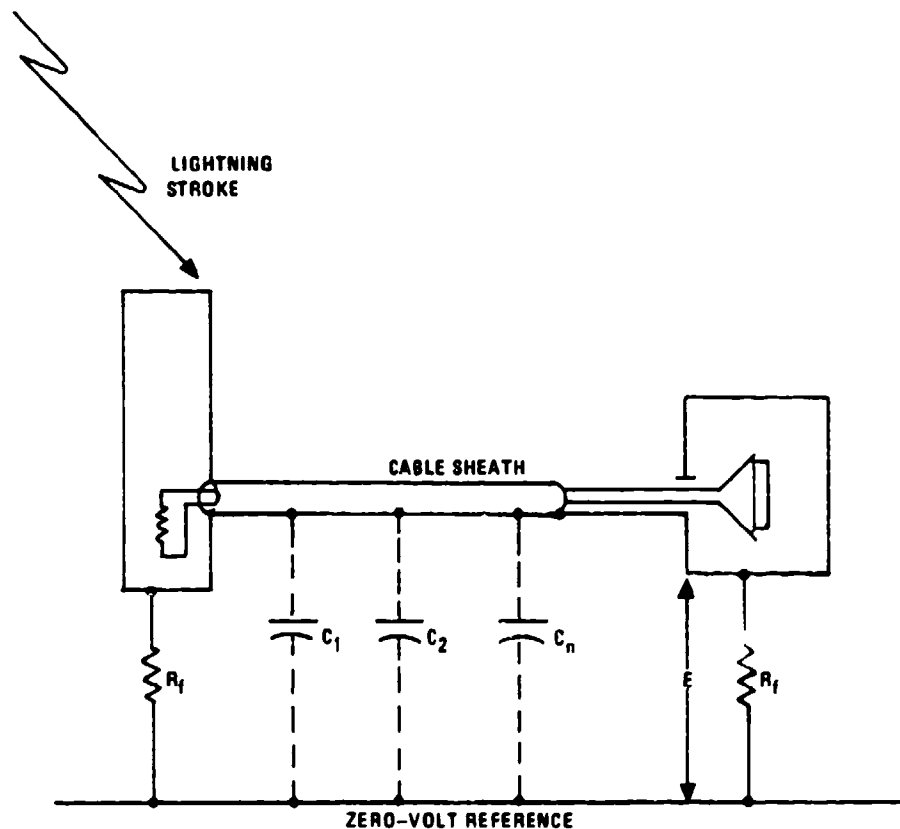


Figure 3-11. Capacitive Coupling of Lightning Energy

This type of lightning threat can be minimized by (1) reducing  $R_{G1}$  to approximately the magnitude of  $R_{G2}$ , (2) the installation of appropriate lightning arresters at the transformer to keep the potential difference between the power conductors and the ground wire and between the primary and secondary windings to within the stress ratings of the transformer, and (3) interconnecting the earth electrode subsystem (to include the water and other utility pipes) with a 1/0 or larger buried copper cable as illustrated by the dotted line in Figure 3-12.

Interconnecting the ground electrodes of the building and transformer pole to form one effective earth contact does not eliminate the lightning threat to the buried cable between the two buildings. As shown, the cable shield is connected to the cabinet, i.e., the building ground. In the event of a lightning strike as shown, Building 1 and its power supply system will be elevated in potential relative to Building 2. In particular, if the distance between the two buildings is more than just a few meters, the inductance, primarily, of the cable shield will prevent the cable from providing the low impedance necessary to keep the two buildings at the same potential. In addition if the shield of the cable is insulated from the earth, as is usually the case, the potential of the cable shield can become high enough with respect to the earth to exceed the breakdown of the insulation.

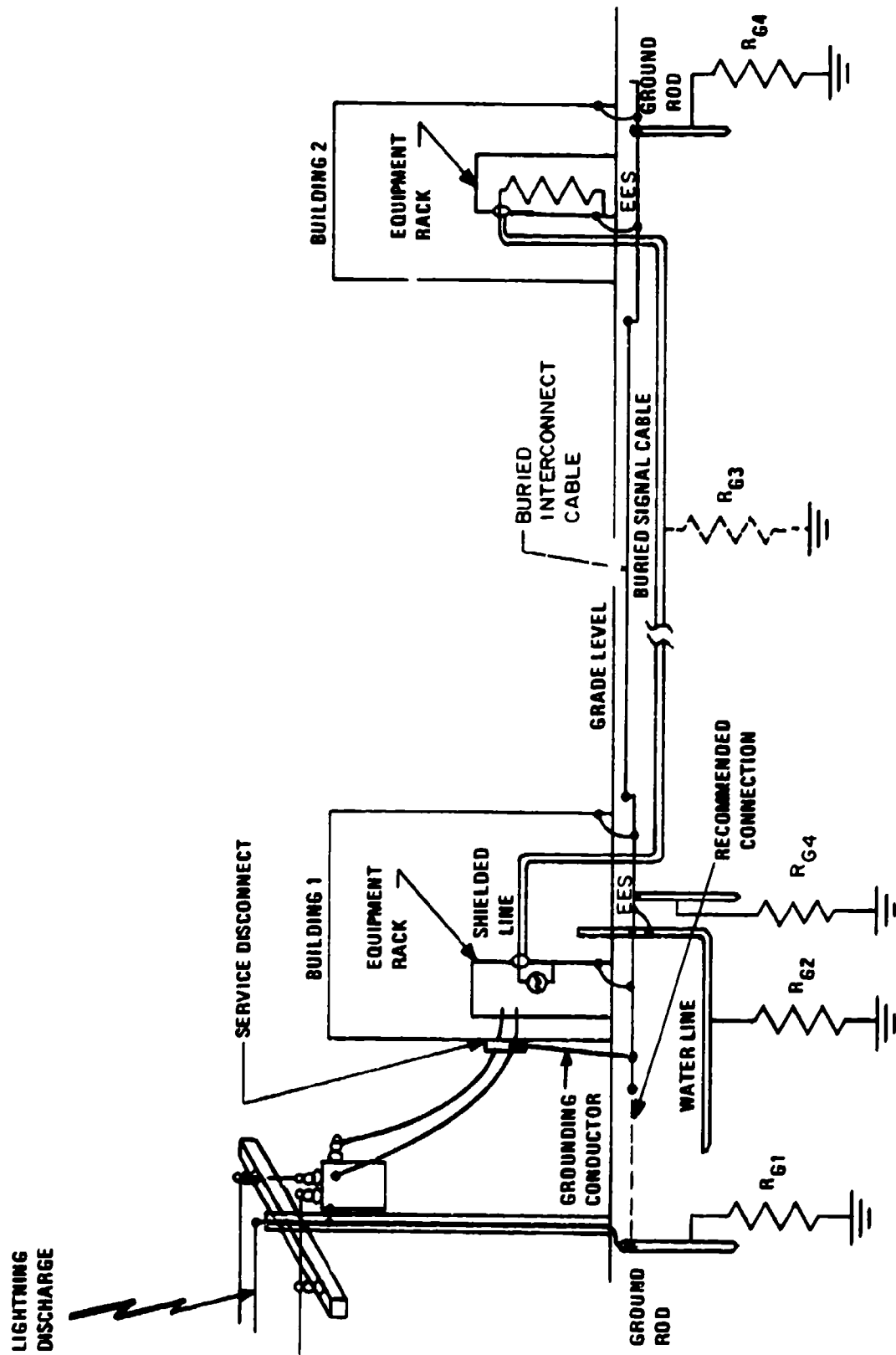


Figure 3-12. Coupling of Lightning Energy Through and Interconnected Facility



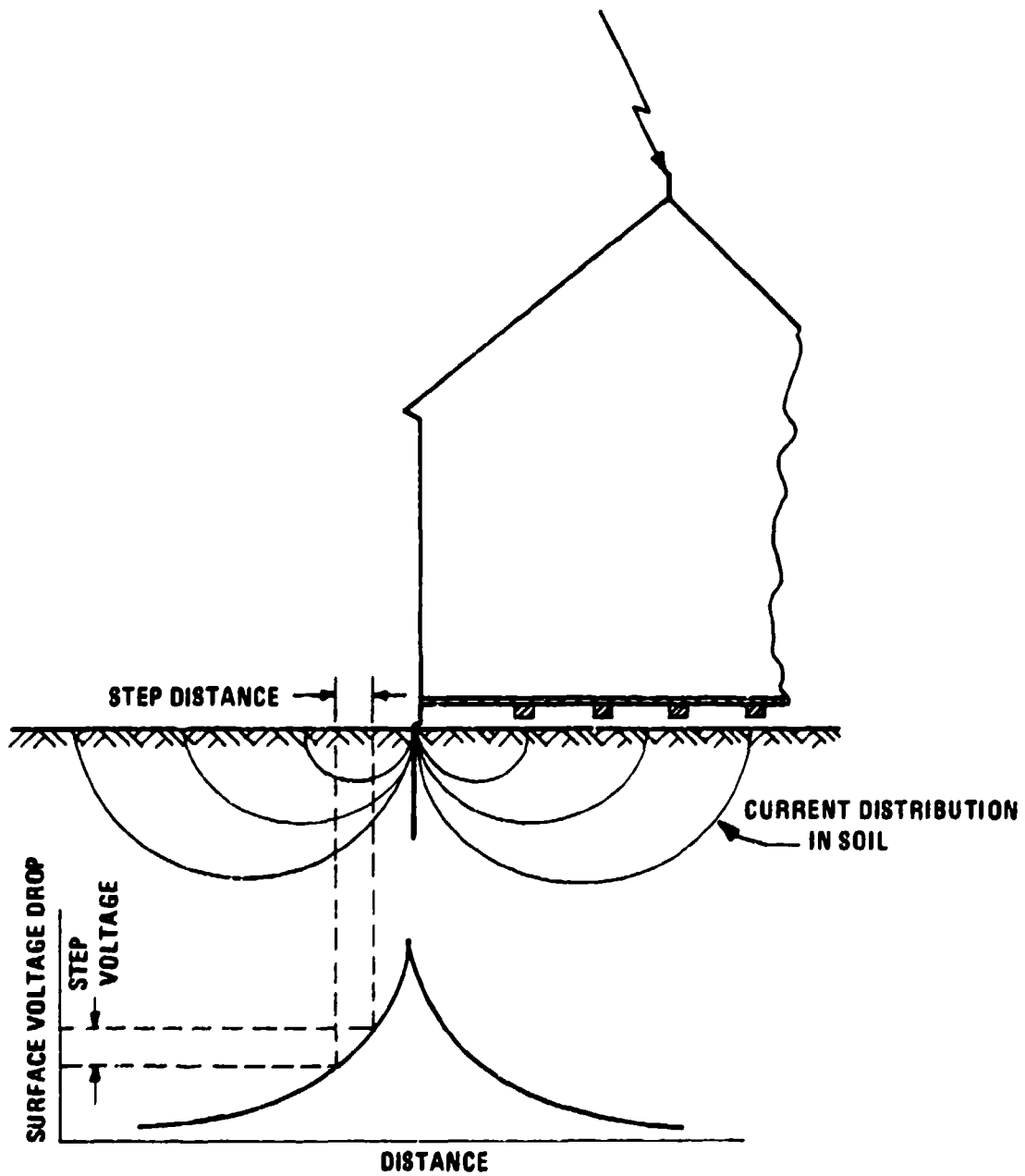


Figure 3-13. Step-Voltage Hazards Caused by Lightning-Induced Voltage Gradients in the Earth

Assume for the moment that Building 1 has an earth electrode subsystem consisting of ground rods interconnected with the cold water system with a net resistance to earth of 3 ohms. With a lightning discharge of 20 kA, the voltage of the complex will rise to 60 kV with respect to Building 2 and that portion of the earth not in the immediate vicinity of Building 1. At Building 1, the cable shield voltage will rise along with that of the building. This voltage pulse will travel down the cable, successively raising the shield potential to as much as 60 kV with respect to the surrounding earth. Such high voltages cause insulation breakdown in the form of tiny pinholes where the lightning energy punches through.

As the lightning pulse travels down the cable, its amplitude diminishes due to cable resistance and dielectric losses. However, the amplitude of the pulse can still be sufficient to damage circuit components in terminating equipment in Building 2. To minimize this damage, surge arresters compatible with the terminating components and hardware should always be provided on such cables. Further information on the use of surge arresters is presented in Volume II, Section 1.3.3.5.

In the event of a lightning stroke, there is a definite personnel hazard posed by the voltage gradient in the soil in the vicinity of the point where the lightning discharge enters the earth. In homogeneous soil, the current rapidly leaves the electrode. The current density is highest near the electrode and rapidly decreases with distance from the electrode. In soil of uniform resistivity, a significant voltage gradient will exist between two points that are differing distances from the electrode. Figure 3-13 illustrates the nature of this voltage variation and shows the hazard encountered by personnel walking (or standing) in the area. The voltage difference across the span of a step can be sufficient to be lethal. As shown earlier, the degree of the hazard is determined by the magnitude of the stroke current, the grounding resistance of the earth electrode, and the distance away from the electrode. No control can be exercised over the current; the threat, however, can be lessened by achieving a low common ground resistance and by minimizing the step potential as discussed in Section 2.8.1.3.

### 3.7 BASIC PROTECTION REQUIREMENTS.

To effectively protect a structure such as a building, mast, tower, or similar self-supporting object from lightning damage, the following requirements must be met:

a. An air terminal of adequate height, mechanical strength and electrical conductivity to withstand the stroke impingement must be provided to intercept the discharge to keep it from penetrating any nonconductive outer coverings of the structure or to prevent it from terminating on antennas, lighting fixtures, transformers, or other devices likely to be damaged or destroyed.

b. A low impedance path from the air terminal to earth must be provided.

c. The resistance of the connection between the discharge path and the earth must be low.

These requirements are met by either (1) an integral system of air terminals, roof conductors, and down conductors, securely interconnected to provide the shortest practicable path to earth, or (2) a separately mounted shielding system such as a metal mast which acts as an air terminal, and a down conductor or an overhead ground wire terminated at the ends (and at intermediate locations, if needed) with down leads connected to earth ground electrodes. Specific design practices are contained in Volume II.

### 3.8 DETERMINING THE NEED FOR PROTECTION.

The degree to which lightning protection is required, is a subjective decision requiring an examination of the relative criticalness of the structure location and its contents to the overall mission of the facility. Those structures containing elements vital to the operational mission such as air traffic control towers, radar installations, navigational aids, and communications centers are examples of facilities which obviously must be protected. However, every building or structure does not require that a lightning protection system be installed. For example, buildings primarily used for the storage of nonflammable materials do not have a critical need for protection.

Three of the factors to consider in ascertaining whether a given structure should have a lightning protection system installed or in determining the relative comprehensiveness of the system are the relative threat of being struck by lightning, the type of construction, and the nature of the facility.

#### 3.8.1 Strike Likelihood.

The relative likelihood of a particular structure being struck by lightning is a function of the keraunic level, i.e., the thunderstorm activity of the locality, the effective height of the structure and its attractive area. Average thunderstorm activity can be determined from the isokeraunic maps shown in Figures 3-2 and 3-3. Then using the techniques described in Section 3.4, estimate the frequency with which strikes to the structure may occur. Use this estimation as one of the inputs to the decision process.

#### 3.8.2 Type of Construction.

Steel frame buildings with metal outer coverings offer the greatest inherent protection against lightning damage. Steel towers also exhibit a high immunity to structural damage. Additional protection for these type buildings will probably be required only for very critical facilities in highly exposed locations. Steel frame buildings with nonconductive, but nonflammable, outer coverings (like brick or other masonry) also offer a high degree of protection against lightning damage. The greatest hazard is posed by pieces of masonry being dislodged by stroke currents passing through the outer coverings to reach the structural steel underneath. Minimal protection consisting of interconnected air terminals to down conductors and steel support columns will be sufficient to prevent this type of structural damage.

Buildings constructed of nonconductive materials such as wood, concrete blocks, or synthetic materials are the most susceptible to destructive damage. A complete auxiliary protection system will be required to prevent lightning damage to buildings utilizing this type of construction.

### 3.8.3 Criticalness to System Mission.

If a strike to the facility poses a threat to human life, either to the occupants of the structure or to those persons whose safety is dependent upon reliable performance of the equipment and people inside the structure, comprehensive lightning protection should be definitely provided even in areas of low thunderstorm activity. At the other extreme, the need for the protection of buildings used primarily to store nonflammable or nonexplosive items is doubtful unless the stored items are critical to system operation, the building is usually exposed, etc. In between these extremes are those structures whose incapacitation would cause an inconvenience or present other difficulties short of life-and-death situations. With these structures, a careful analysis must be made to determine the relative likelihood of outages from lightning in comparison to normal equipment failures, downtime for maintenance, and other routine occurrences.

Though not directly related to the protection of electrical or electronic installations, Reference 3-10 is recommended for further guidance in performing the tradeoff analyses to determine the degree of lightning protection required for specific facilities.

### 3.9 APPLICABLE CODES.

The Lightning Protection Code, NFPA No. 78, issued by the National Fire Protection Association (3-9) contains the basic requirements for the minimization of personnel hazards in the event of a lightning strike to the structure.

The requirements of NFPA No. 78, however, are not sufficient to protect the electrical distribution system, signal and control cables, or sensitive electronic equipment from surges produced by either direct or indirect strokes. Thus additional steps such as providing lightning arresters on power lines and on outside signal and control cables, providing counterpoise cables for overhead and underground cables, providing comprehensive electromagnetic shielding on sensitive cables, and installing fast response surge protection devices on circuits exposed to lightning discharges should be taken. MIL-STD-188-124A refers.

3.10 REFERENCES.

- 3-1. M.A. Uman, Lightning, McGraw-Hill, New York NY (1969).
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## CHAPTER 4

## FAULT PROTECTION SUBSYSTEM

**4.1 FAULT PROTECTION.** For effective fault protection, a low resistance path must be provided between the location of the fault and the transformer supplying the faulted line. The resistance of the path must be low enough to cause ample fault current to flow and rapidly trip breakers or blow fuses. The necessary low resistance return path inside a building is provided by the grounding (green wire) conductor and the interconnected facility ground network. An inadvertent contact between energized conductors and any conducting object connected to the grounding (green wire) conductor will immediately trip breakers or blow fuses. In a building containing a properly installed third-wire grounding network, as prescribed by MIL-STD-188-124A, faults internal to the building are rapidly cleared regardless of the resistance of the earth connection.

**4.1.1 Power System Faults.**

A power system fault is either a direct short or an arc (continuous or intermittent) in a power distribution system or its associated electrical equipment. These faults are hazardous to personnel for several reasons:

- a. Fault currents flowing in the ground system may cause the chassis of grounded equipment to be at a hazardous potential above ground.
- b. The energy in a fault arc can be sufficient to vaporize copper, aluminum, or steel. The heat can present a severe burn hazard to personnel.
- c. There is a fire hazard associated with any short circuit or arc.
- d. Burning insulation can be particularly hazardous because of the extremely toxic vapors and smoke which may be produced.

Some common causes of electrical system faults are:

- a. Rodents getting between ground and phase conductors.
- b. Water infiltration.
- c. Moisture in combination with dirt on insulator surfaces.
- d. Breakdown of insulation caused by thermal cycling produced by overloads.
- e. Environmental contaminants.
- f. Damage during installation.
- g. System age deterioration.

Figure 4-1 illustrates how personnel hazards are developed by improper installation and fault conditions. Suppose that one phase of the 230-volt line accidentally contacts the motor frame. If the motor is not grounded, its frame will rise to 133 volts, and anyone coming in contact with it would be subject to a lethal shock if simultaneous contact is made with a grounded object. To prevent this situation from arising, the motor frame must be grounded via the green wire. The resistance of the fault path must be low enough to permit the fault current to trip the overload protector and interrupt the fault. If the resistance of the fault path is too large, the fault current will not be enough to trip the overload protectors. Thus to minimize both shock and fire hazards, the resistance of the fault path must be as low as possible. However, the fault protection subsystem normally does not depend on the earth electrode subsystem to trip overcurrent devices. The fault current normally flows through the green wire (grounding conductor) to the source side of the first service disconnect means where the green wire and the neutral are tied together. The fault current then flows through the neutral to the transformer to complete the circuit. This path functions completely independent of the connection to the earth electrode subsystem.

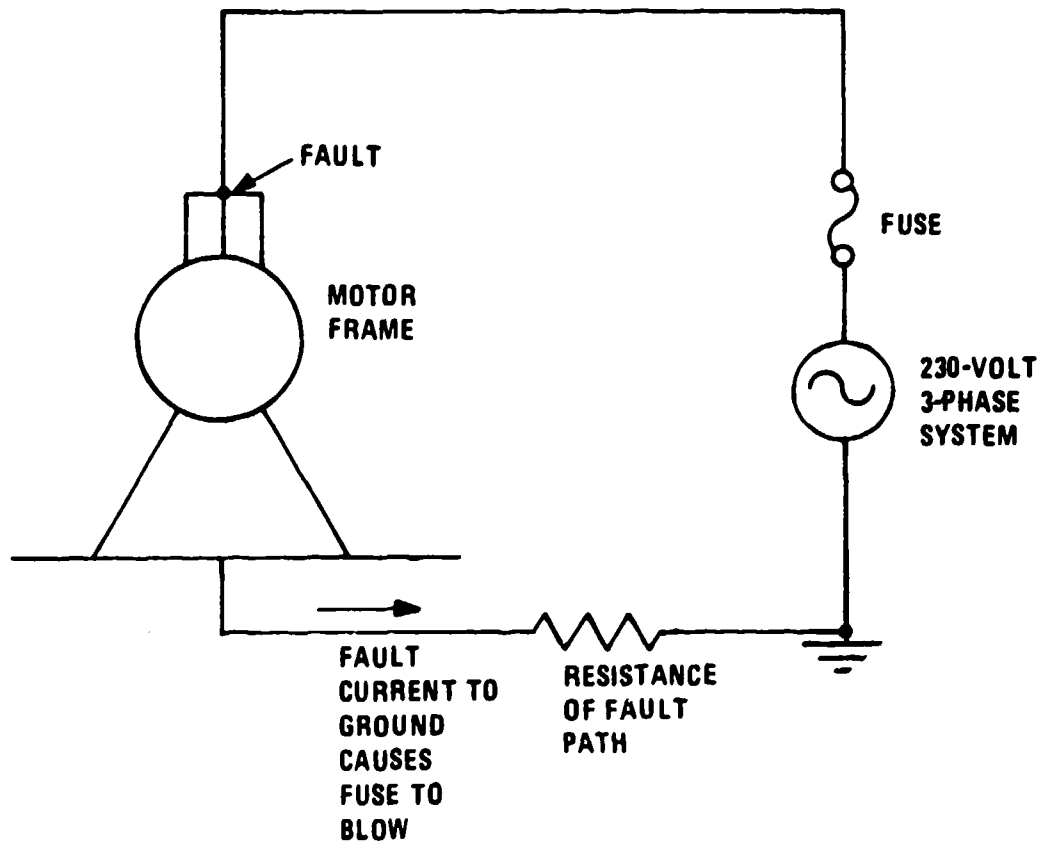


Figure 4-1. Grounding for Fault Protection

Fault clearance in power distribution systems is normally provided by circuit breakers, fuses, or overload relays in each phase. These devices provide personnel protection only if the fault current is sufficient to trip the over-current device. They generally however do not have response times which are adequate to protect the individual if he happens to be in direct contact with the energized object.

**4.1.2 Ground-Fault-Circuit-Interrupter (GFCI).** High resistance faults (low and moderate currents of 5 milliamperes or more) can be cleared rapidly with a device called a ground-fault-circuit-interrupter (GFCI). The GFCI contains an electronic circuit which continuously monitors the difference between the current supplied to the load and the current returned from the load. If this difference is not zero, some current must be leaking to ground. When this leakage current exceeds a preset value, the GFCI will act to interrupt the power to the circuit. GFCI's are so sensitive that they can be set to interrupt power fault currents as low as 2 milliamperes. Experiments with dogs have shown that trip currents of 5 milliamperes or less will prevent electrocution. (GFCI's have proven so effective as protection against electric shock that the National Electrical Code requires that all 15 and 20 ampere bathroom, garage, and outdoor receptacles in family dwelling units and in circuits set up at construction sites be protected with a GFCI. MIL-STD-188-124A also recommends they be installed on 120 volt single phase 15 and 20 ampere receptacles of C-E facilities.)

## **4.2 EARTH CONNECTION.**

Historically, grounding requirements arose from the need to protect personnel, equipment, and facilities from lightning strokes and from industrially generated static electricity. Structures, as well as electrical equipment, were connected to earth, i.e., grounded, to provide the path necessary for lightning and static discharges. As utility power systems developed, grounding to earth was found to be necessary for safety. All major components of the system such as generating stations, substations, and distribution systems are earth grounded to provide a path back to the generator for the fault currents in case of transmission line trouble. The path to earth should have as low a resistance as possible. A low resistance minimizes the potential difference between equipments connected to the earth electrode subsystem when fault currents flow. Thus personnel who come in contact with two or more pieces of equipment at one time are protected.

Ideally, the earth connection should exhibit zero resistance between the earth and the equipment and facilities connected to it. Any physically realizable connection, however, will exhibit a finite resistance to earth. The economics of the design of the earth electrode subsystem involves a trade-off between the expense necessary to achieve a low resistance and the satisfaction of minimum subsystem requirements. The 10 ohm design objective of MIL-STD-188-124A is considered such a trade-off.

## **4.3 AC POWER LINE GROUND.**

The grounding conductor (green wire) in a single-phase 115/230 volt ac power distribution system in a facility is one of four leads, the other three being the two phase or "hot" leads (black/red) and the neutral lead (white wire). The green wire is a safety conductor designed to carry current only in the event of a fault. The "hot" leads are connected from the first service disconnect to the high sides of the secondary of the distribution transformer and the neutral is connected to the center tap which is grounded to a ground terminal at the transformer. When a single transformer supplies power to only one communications building, for fault



protection the grounding conductor shall be grounded on the source side of the first service disconnect to the earth electrode subsystem and also to the ground terminal at the distribution transformer. For 3-phase wye systems a five-wire service entry cable consisting of one neutral, one grounding, and three phase conductors shall be employed. In either case, when a single transformer supplies power to a single building, the safety ground (green wire) shall be grounded to the earth electrode subsystem at the supply side of the first service disconnect of the facility as well as at the distribution transformer as shown in Figure 4-2. The neutral shall also be grounded at both locations.

When a single transformer supplies power to more than one C-E building and if noise or hum is encountered in C-E circuits or equipments, the neutral should be lifted or removed from ground at each service disconnect. In this case the neutrals from each building are grounded at the distribution transformer only (see Figure 4-3).

To protect personnel from exposure to hazardous voltages, all exposed metallic elements of electrical and electronic equipment are connected to ground with the green wire. Then, in the event of inadvertent contact between the hot lead and chassis, frame, or cabinet through human error, insulation failure, or component failure, a direct fault clearance path is established to quickly remove the hazard.

Grounding of a 3-phase wye power distribution system is accomplished similarly to the single phase system. The connections for a typical system are shown in Figure 4-3. As in single phase systems, the neutral lead is bonded to the green wire at the supply side of the first service disconnecting means and grounded to the earth electrode subsystem as well as to the ground terminal at the distribution transformer. If one transformer supplies power to more than one C-E building, the neutral is lifted from ground at the service disconnect.

A 3-phase system served by a transformer with a delta connected secondary will require the use of a grounding transformer to ground the system and establish a neutral. The grounding transformer may be either a "zig-zag" or "wye-delta" type, both of which have leads which are attached to each of three phases and a fourth lead which is grounded and serves as the neutral. The typical connections for a grounding transformer are shown in Figure 4-4.

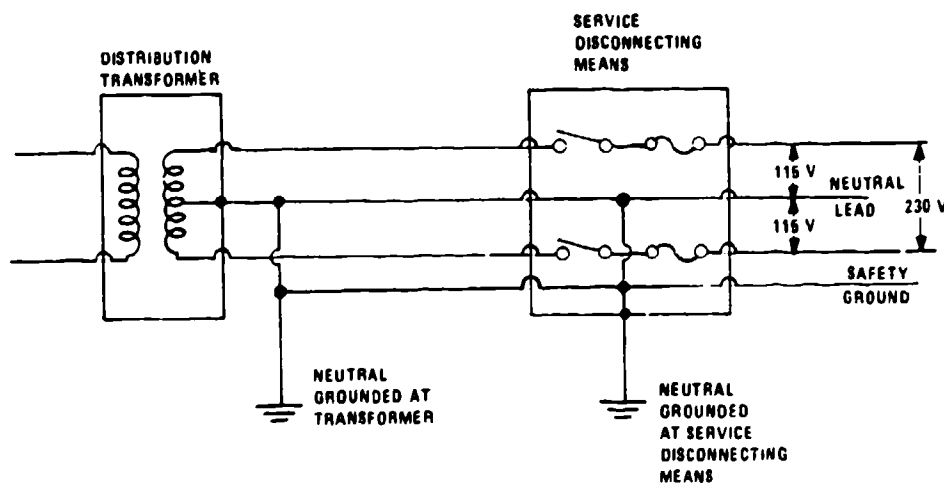
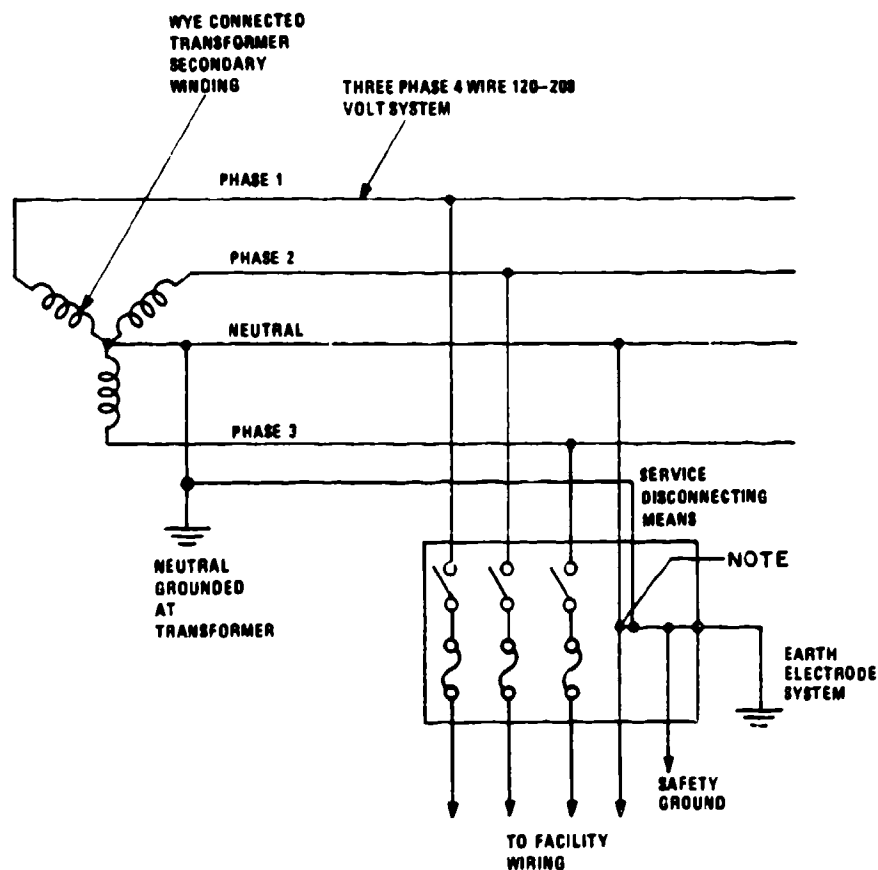


Figure 4-2. Single-Phase 115/230 Volt AC Power Ground Connections

**4.4 TEST EQUIPMENT.** Test equipments are available to measure the resistances and impedances of the fault protection subsystems including the grounding (green) conductor as well as the signal reference subsystem (equipotential plane) which may at times become part of the fault protection subsystem. These equipments can measure the impedances (at 60 Hz) of each path from the equipment having the fault to the first service disconnect means and therefore assist in determining the value of the fault current over each path. The information will in turn be beneficial in determining or predicting the degree of interference which may be anticipated should a fault current be superimposed on the signal reference subsystem. (4-1 and 4-2)



NOTE: Lift when single transformer supplies power to more than one building or because of objectionable current, noise or interference.

Figure 4-3. Three-Phase 120/208 Volt AC Power System Ground Connections

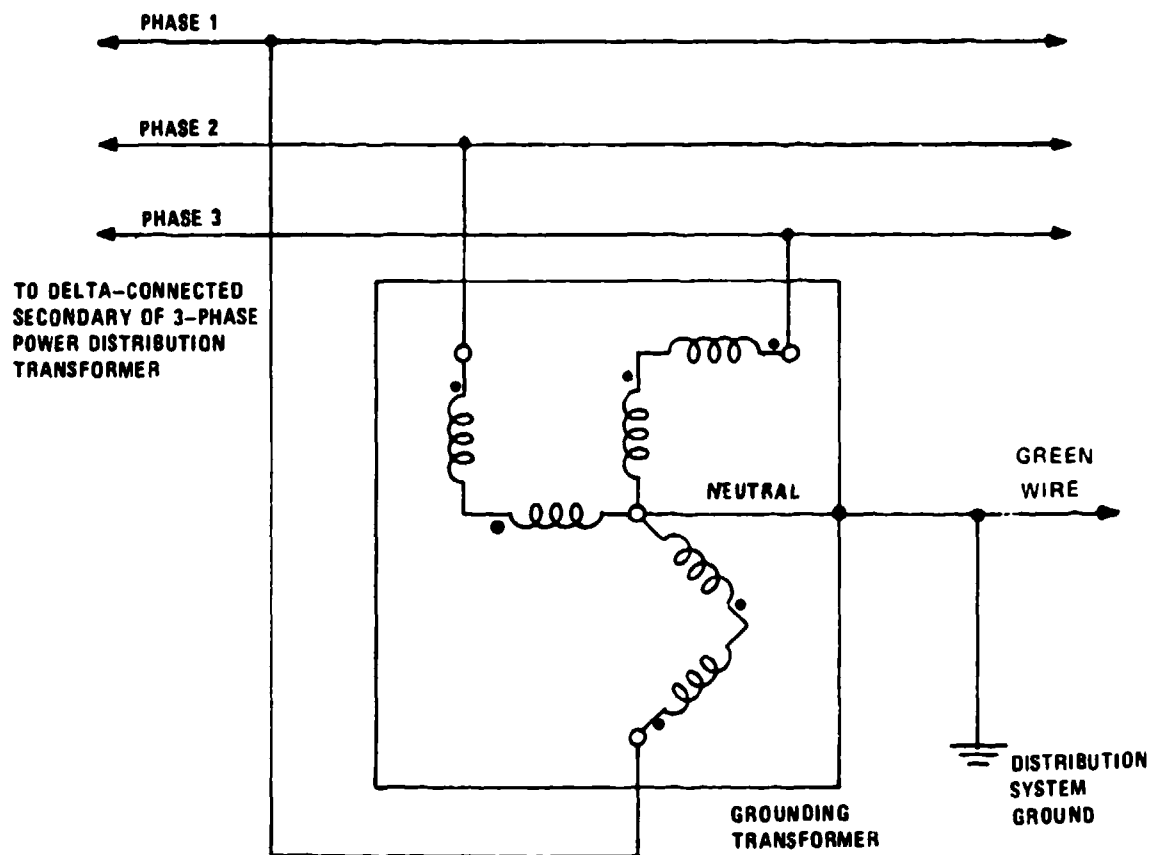


Figure 4-4. Connections for a Three-Phase "Zig-Zag" Grounding Transformer

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## CHAPTER 5

## GROUNDING OF SIGNAL REFERENCE SUBSYSTEM

5.1 INTRODUCTION.

Signal circuits are grounded and referenced to ground to (1) establish signal return paths between a source and a load, (2) control static charge, or (3) provide fault protection. The desired goal is to accomplish each of these three grounding functions in a manner that minimizes interference and noise.

If a truly zero impedance ground reference plane or bus could be realized, it could be utilized as the return path for all currents -- power, control, audio and rf -- present within a system or complex. This ground reference would simultaneously provide the necessary fault protection, static discharge, and signal returns. The closest approximation to this ideal ground would be an extremely large sheet of a good conductor such as copper, aluminum, or silver underlying the entire facility with large risers extending up to individual equipments. The impedance of this network at the frequency of the signal being referenced is a function of conductor length, resistance, inductance, and capacitance. When designing a ground system in which rf must be considered, transmission line theory must be utilized.

5.2 CONDUCTOR CONSIDERATIONS.5.2.1 Direct Current Resistance.

The resistance,  $R_{dc}$ , of a conductor of uniform cross section is proportional to the length and inversely proportional to the cross-sectional area, that is

$$R_{dc} = \rho l / A \text{ ohms} , \quad (5-1)$$

where  $\rho$  is the resistivity of the conductor material,  $l$  is the length of the conductor in the direction of current flow, and  $A$  is the cross-sectional area of the conductor. Values of  $R_{dc}$  for the standard sizes of wire and cable are given in Table 5-1. (For data on wire sizes not shown in this table, consult References 5-1 and 5-2.)

At dc, the resistance of the conductor is the controlling factor. Except for very unusual situations (such as when the signal to be processed is very low in amplitude or where the interfacing equipments are very far apart physically), an adequate ground can generally be realized for dc in a relatively economical manner utilizing low resistivity materials such as copper and aluminum. Most systems, however, employ other than dc signals. Therefore, the frequency-dependent properties of the conductors become important.

5.2.2 Alternating Current Impedance. The ac impedance of a conductor is composed of two parts: the ac resistance and the reactance. Both the ac resistance and the reactance of a conductor vary with frequency as a result of skin effect.

Table 5-1

## Properties of Annealed Copper Wire

AWG No.	Diameter		Cross-Sectional Area		Resistance in Ohms	
	mils	mm	cmil	mm <sup>2</sup>	per 1000 ft	per km
4/0	460.0	11.7	211 600	107.2	0.049	0.161
3/0	409.6	10.4	167 800	85.0	0.062	0.203
2/0	364.8	9.3	133 100	67.4	0.078	0.256
1/0	324.9	8.3	105 500	53.4	0.098	0.322
1	289.3	7.3	83 690	42.4	0.124	0.407
2	257.6	6.5	66 370	33.6	0.156	0.512
4	204.3	5.2	41 740	21.1	0.248	0.814
6	162.0	4.1	26 250	13.3	0.395	1.296
8	128.5	3.3	16 510	8.4	0.628	2.060
10	101.9	2.6	10 380	5.3	0.999	3.278
12	80.8	2.1	6 530	3.3	1.588	5.210
14	64.1	1.6	4 107	2.1	2.525	8.284
16	50.8	1.3	2 583	1.3	4.016	13.176
18	40.3	1.0	1 624	0.8	6.385	20.948
20	31.9	0.8	1 022	0.5	10.150	33.300

### 5.2.2.1 Skin Effect.

Whereas a direct current is uniformly distributed over the cross-sectional area of a conductor, alternating current tends to concentrate near the surface of the conductor. The higher the frequency, the greater the concentration near the surface. This physical phenomenon is called skin effect. A measure of the degree of penetration of the currents into the conductor is given by the skin depth,  $\delta$ .  $\delta$  is defined as the depth at which the current density is attenuated to  $1/e = 1/2.718 \approx 0.37$  of its value at the conductor surface. Skin depth may also be interpreted as the equivalent thickness of a hollow conductor carrying a uniform distribution over its cross-sectional area, having the same external shape as the actual conductor, and having a dc resistance exactly the same as the ac resistance of the conductor.

For conductors whose thickness is at least three times the skin depth, this depth is given by (5-3).

$$\delta = 5033 \sqrt{\frac{\rho}{\mu_r f}} \text{ cm,} \quad (5-2)$$

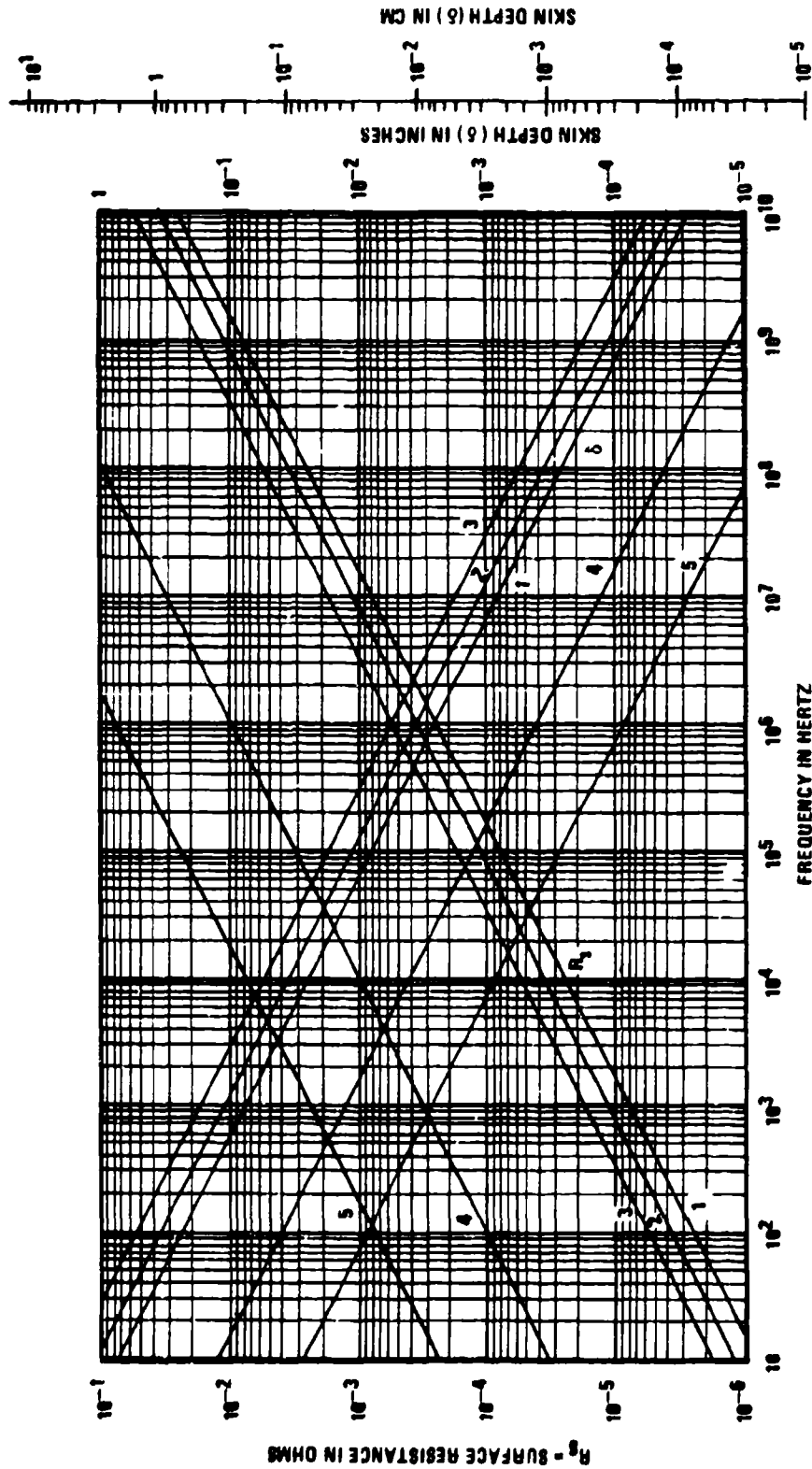
where  $\rho$  is the resistivity of the material in ohm-cm,  $f$  is the frequency in hertz, and  $\mu_r$  is the relative permeability of the material. The skin depth for various metals is given in Table 5-2 and Figure 5-1. Note that copper has a skin depth of 0.34 inch (8.63 mm) at 60 Hz but only .00026 inch (0.066 mm) at 1 MHz.

Table 5-2

Parameters of Conductor Materials (5-4)

Material*	$\rho$ ( $\Omega$ -cm)	$\delta$ (cm)	$R_s$ ( $\Omega$ )
Silver	$1.62 \times 10^{-6}$	$6.41/\sqrt{f}$	$2.52 \times 10^{-7} \sqrt{f}$
Copper	$1.73 \times 10^{-6}$	$6.62/\sqrt{f}$	$2.61 \times 10^{-7} \sqrt{f}$
Aluminum	$2.69 \times 10^{-6}$	$8.25/\sqrt{f}$	$3.26 \times 10^{-7} \sqrt{f}$
Brass	$6.37 \times 10^{-6}$	$12.70/\sqrt{f}$	$5.01 \times 10^{-7} \sqrt{f}$
Solder	$14.2 \times 10^{-6}$	$18.96/\sqrt{f}$	$7.48 \times 10^{-7} \sqrt{f}$

\*  $\mu_r = 1$



- (1) COPPER
- (2) ALUMINUM
- (3) BRASS
- (4) SOFT IRON
- (5) PERMALLOY 78

Figure 5-1. Surface Resistance and Skin Depth for Common Metals (5-5)

5.2.2.2 AC Resistance.

The ac resistance of a conductor of any shape can be determined from the skin depth if both the thickness and the radius of curvature of the conductor are much greater than the skin depth and if the radius of curvature does not vary too rapidly around the conductor's perimeter. For a conductor meeting these conditions, the ac resistance per unit length is

$$R_{ac} = \frac{\rho}{\delta P} \quad \text{ohms/meter, or} \quad (5-3)$$

$$= \frac{R_s}{P} \quad \text{ohms/meter} \quad (5-4)$$

where  $P$  is the circumference of the conductor and  $R_s$  is the surface resistance of the conductor. The surface resistance is defined as the ac resistance of a surface of equal length and width and is given by

$$R_s = \frac{\rho}{\delta} = 1.987 \times 10^{-4} \sqrt{\mu_r f \rho} \quad (5-5)$$

The surface resistance for various metals is also shown in Figure 5-1 and Table 5-2.

The ratio of the ac resistance to the dc resistance is called the resistance ratio of a conductor. Skin effect causes the resistance ratio to be greater than unity. The resistance ratio for straight cylindrical wires is given in Figure 5-2 in terms of a parameter  $X$  defined as

$$X = \sqrt{\frac{8\pi\mu_r f}{R_{dc} \times 10^9}} \quad (5-6)$$

where  $\mu_r$  is the relative permeability of the conductor,  $f$  is the frequency in hertz, and  $R_{dc}$  is the dc resistance in ohms for 1 cm of conductor.\* In the case of copper wire, Equation 5-6 becomes

$$X_{cu} = 2.71 \times 10^{-4} d_m \sqrt{f} \quad (5-7)$$

where  $d_m$  is the wire diameter in mils, or becomes

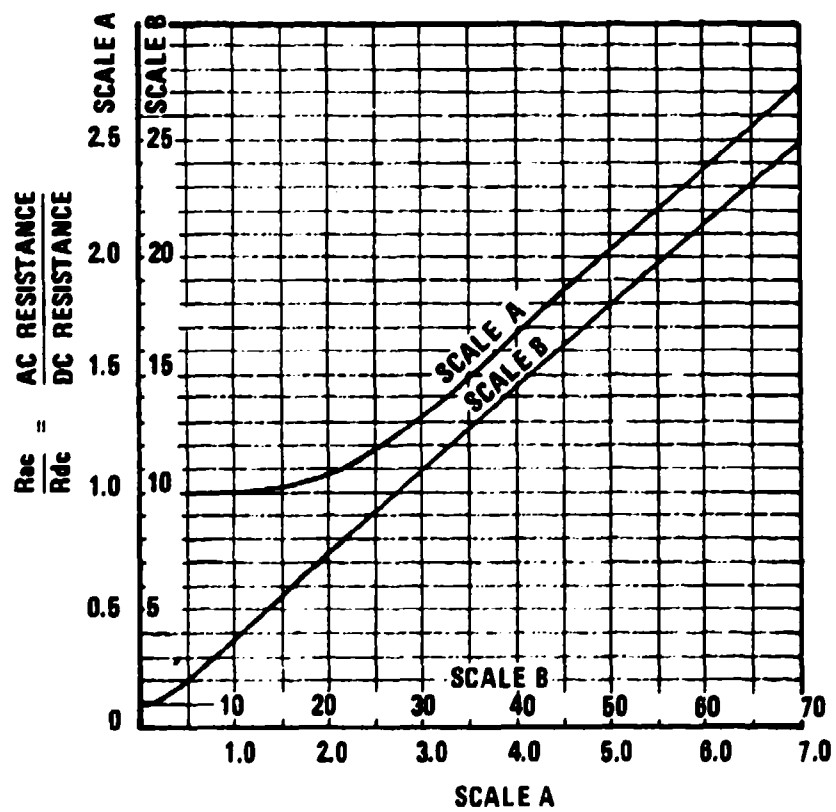
$$X_{cu} = 1.07 \times 10^{-2} d_m \sqrt{f} \quad (5-8)$$

where  $d_m$  is diameter in mm.

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\*It should be noted that Equation 5-6 applies at all frequencies, whereas Equations 5-3 and 5-4 apply only under the conditions stated.





$$X = \sqrt{\frac{8\pi\mu_r f}{R_{dc} \cdot 10^9}}$$

Figure 5-2. Resistance Ratio of Isolated Round Wires (5-6)

### 5.2.2.3 Reactance.

The reactance of the conductor is generally inductive and is given by the product of the radian frequency,  $\omega$ , and the self-inductance,  $L$ , of the conductor. The self-inductance of a conductor is a measure of that property which causes an opposition to a change in the current flowing in the conductor. Because skin effect redistributes the current within a conductor with changes in frequency, the inductance of the conductor does vary with frequency.

The self-inductance of a straight round wire is given (5-6) by

$$L = 0.00508\ell \left( 2.303 \log \frac{4\ell}{d} - 1 + \mu_r \kappa \right) \mu\text{H} \quad (5-9)$$

where  $\ell$  is the length in inches,  $d$  is the diameter in inches, and  $\kappa$  is a skin effect correction factor which may be determined (for copper) from Figure 5-3. For  $\ell$  and  $d$  in centimeters, Equation 5-9 becomes

$$L = 0.002\ell \left( 2.303 \log \frac{4\ell}{d} - 1 + \mu_r \kappa \right) \mu\text{H} \quad (5-10)$$

For materials other than copper,  $\kappa$  can be obtained from Figure 5-3 by using  $f' = f(\rho_c/\rho)$  instead of the actual frequency  $f$ , where  $\rho$  is the resistivity of the material and  $\rho_c$  is the resistivity of copper. For low frequencies where the current flow can be assumed to be uniform across the conductor cross-section, the inductance of a round straight wire of length  $\ell$ , diameter  $d$ , and relative permeability  $\mu_r$  (if surrounded by air) is

$$L_{LF} = 0.00508\ell \left( 2.303 \log \frac{4\ell}{d} - 1 + \frac{\mu_r}{4} \right) \mu\text{H} \quad (5-11)$$

where all the dimensions are in inches. As the frequency increases, a limiting value of inductance,  $L_{HF}$ , is approached:

$$L_{HF} = 0.00508\ell \left( 2.303 \log \frac{4\ell}{d} - 1 \right) \mu\text{H} \quad (5-12)$$

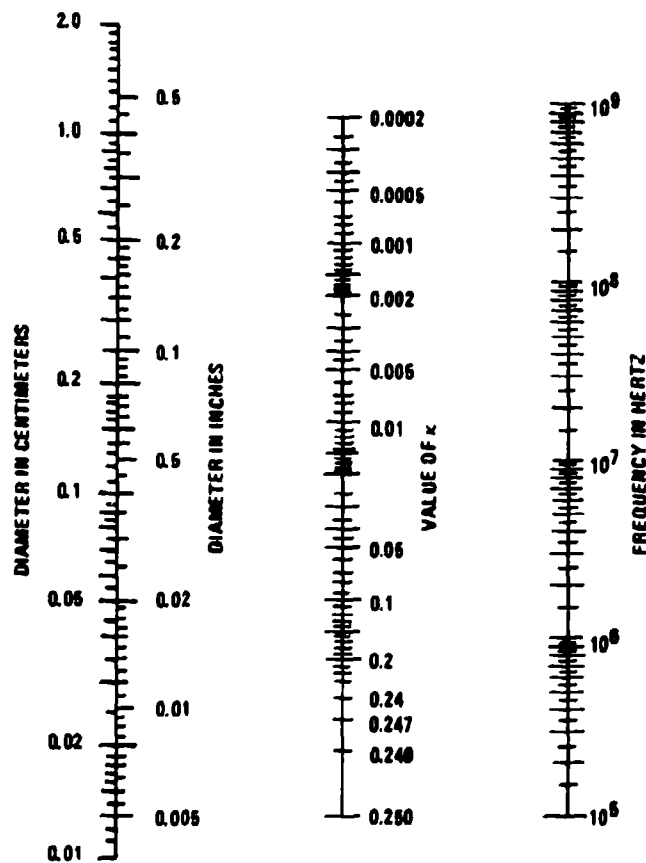


Figure 5-3. Nomograph for the Determination of Skin Effect Correction Factor (5-6)

In Equations 5-11 and 5-12, the constant 0.00508 becomes 0.002 when  $l$  and  $d$  are in cm.

Figure 5-4 gives the value of  $L_{LP}$  for a 1/0 AWG solid round copper conductor as a function of length, and  $L_{HP}$  for various wire lengths and diameters is given in Figure 5-5.

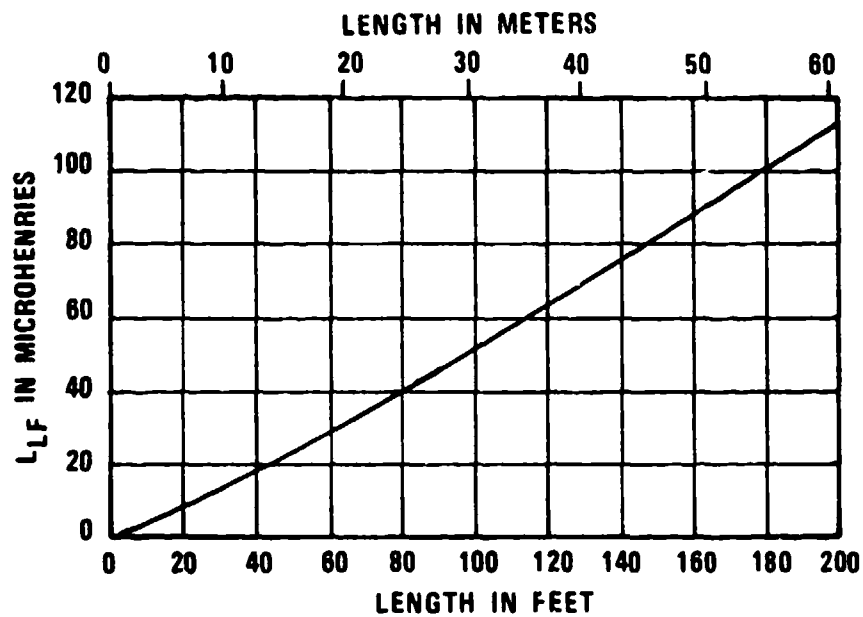


Figure 5-4. Low Frequency Self-Inductance versus Length for 1/0 AWG Straight Copper Wire (5-7)

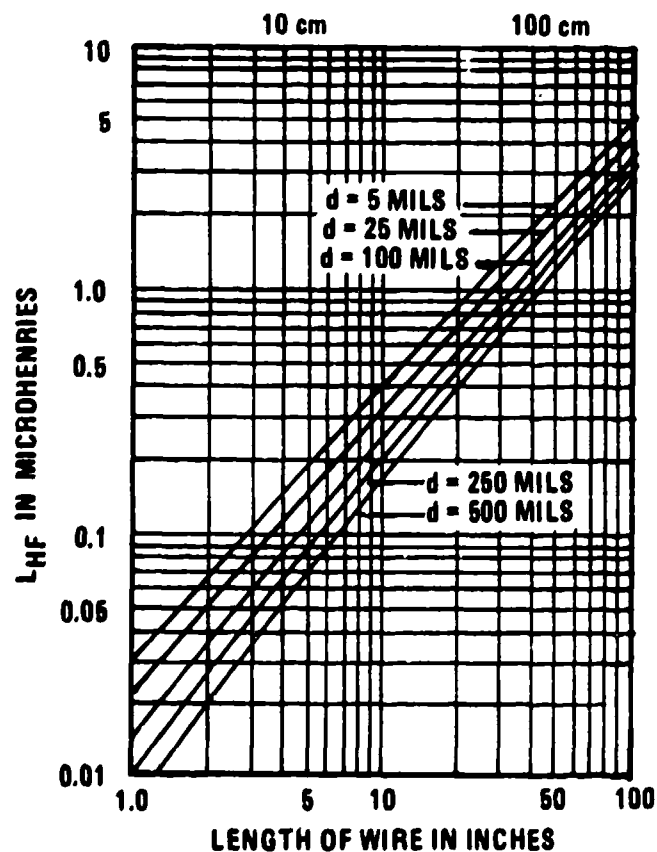


Figure 5-5. Self-Inductance of Straight Round Wire at High Frequencies (5-6)

**5.2.2.4 Proximity Effect.** When two or more conductors are in close proximity, the current flowing in one conductor is redistributed because of the magnetic field produced by the current in the other conductor. This effect is an extension of skin effect and is called proximity effect. The proximity effect tends to increase the ac resistance of a conductor to a value greater than that due to simple skin effect.

### **5.2.3 Resistance Properties vs Impedance Properties.**

Although skin effect exists at all frequencies, it becomes more significant as the frequency increases. The reactance of a conductor also increases with frequency to further increase the conductor impedance above its dc value. To design an effective ground system one must consider the relative effects of the dc resistance, the ac resistance, and the inductance upon the total impedance of a ground conductor.

Using Equation 5-1, the dc resistance of round wire conductors can be calculated. The dc resistance per 1000 feet for four standard size copper cables is given in Table 5-3. Table 5-4 gives the dc resistance and (for 60 Hz) the ac resistance, the inductance and the total impedance of various size and length conductors as determined from Table 5-3 and from Equation 5-12. At a frequency of 1 MHz, these same characteristics for 30-meter (100-foot) lengths are given in Table 5-5 as calculated from Equations 5-3 and 5-12. Note that for the larger wires (No. 2 AWG or larger) the inductance of the long (> 100 feet) cables determines the magnitude of the impedance. Also note that for the same length cables there is not as much difference in the impedance magnitudes of a small and a large cable as there is in the resistance of the two cable sizes. For example, the ratio of the dc resistance of a 30-meter (100-foot) length of No. 12 AWG copper cable to the dc resistance of a 30 meter (100 feet) of 1/0 AWG copper cable is  $0.15880/0.0098 = 16.20$ . Since the ac resistance at 60 Hz is approximately the same as the dc resistance, the ratio of the 60 Hz ac resistance of the two cables is also 16.20. At a frequency of 1 MHz the ratio of the ac resistance becomes  $1.23/0.307 = 4.01$ . However, the 60 Hz impedance ratio is only  $0.1605/0.0226 = 7.10$  and the 1 MHz impedance ratio is only  $382.65/329.49 = 1.16$ . These ratios are tabulated in Table 5-6 for comparison. From Tables 5-3 through 5-6 and the above example, the following conclusions are made:

- a. Because of the inductance, the advantages offered by a large cable such as 1/0 AWG are less than they might appear to be from a comparison of the dc resistance values.
- b. The advantage offered by a large cable, e.g., 1/0 AWG, will be somewhat more pronounced for relatively short conductor lengths than for long conductor runs. This is true because inductance increases more rapidly with length than does resistance (see Equations 5-1 and 5-9).
- c. Because of the lack of dramatic improvement in ac impedance of large cables over smaller cable sizes for long runs, consideration of materials and labor costs are relatively important and may be the deciding factor.
- d. Since even 1/0 AWG cables exhibit impedances from 22.6 mΩ to 115.8 mΩ for lengths of 30 meters (100 feet) and 137 meters (450 feet), respectively, the control of stray currents should be an essential objective in any signal grounding system.

Table 5-3

## DC Parameters of Some Standard Cables

<u>Size</u> (AWG)	<u>Diameter</u> (mils)	<u>DC Resistance</u> (Ohms/1000 ft)
No. 12	80.81	1.588
No. 8	128.5	0.6282
No. 2	257.6	0.1563
1/0	324.9	0.09827

Table 5-4

## Sixty-Hertz Characteristics of Standard Cables

<u>Size</u> (AWG)	<u>Length</u> (Ft)	<u>R<sub>ac</sub></u> (Ω)	<u>L</u> (μH)	<u>X<sub>L</sub></u> (Ω)	<u>Z</u> (Ω)
No. 12	30	0.04764	16.532	0.00623	0.0480
No. 12	100	0.15880	62.447	0.02354	0.1605
No. 8	30	0.01885	15.684	0.00591	0.0197
No. 2	30	0.00469	14.411	0.00543	0.0072
No. 2	100	0.01563	55.379	0.02088	0.0261
No. 2	150	0.02344	86.777	0.03271	0.0402
1/0	30	0.00294	13.987	0.00527	0.0060
1/0	100	0.00980	53.964	0.0226	0.0060
1/0	150	0.01470	84.654	0.03191	0.0351
1/0	300	0.02940	181.987	0.08861	0.0746
1/0	450	0.04410	284.105	0.10710	0.1158

Table 5-5

## One-Megahertz Characteristics of Standard Cables

<u>Size</u> AWG	<u>Length</u> (Ft)	<u>R<sub>dc</sub></u> (Ω)	<u>R<sub>ac</sub></u> (Ω)	<u>L</u> (μH)	<u>X<sub>L</sub></u> (Ω)	<u>Z</u> (Ω)
No. 12	100	0.1588	1.23	60.9	382.65	382.65
No. 2	100	0.0156	0.387	53.8	338.03	338.03
1/0	100	0.0098	0.307	52.44	329.49	329.49

Table 5-6

## Impedance Comparisons Between No. 12 AWG and 1/0 AWG

<u>Frequency</u>	<u>Length</u>	$\frac{R_{dc} \text{ (No. 12 AWG)}}{R_{dc} \text{ (1/0 AWG)}}$	$\frac{R_{ac} \text{ (No. 12 AWG)}}{R_{ac} \text{ (1/0 AWG)}}$	$\frac{Z \text{ (No. 12 AWG)}}{Z \text{ (1/0 AWG)}}$
60 Hz	30 ft.	16.20	16.20	9.23
60 Hz	100 ft.	16.20	16.20	7.99
1 MHz	100 ft.	16.20	4.01	1.16

5.2.4 Effects of Geometry. Many conductor shapes can be used in the signal ground network. As is the case for the solid round conductor, the impedance of other configured conductors is dependent upon the current distribution in the conductor and hence upon the signal frequency.

#### 5.2.4.1 Stranded Cable.

A stranded cable consists of a number of wires in close proximity twisted about each other; it is more flexible than a solid conductor of the same cross-sectional area. Because of the close proximity of the wires, the skin effect within the cable redistributes most of the current to outer wires. These outer wires are in the form of a coil (due to the lay of the strand), thus increasing the self-inductance of the cable. Skin effect also increases the ac resistance as the frequency is increased.

For a given cable size, both the ac resistance and the self-inductance of a stranded conductor are greater than those of a solid round conductor. Because of their ineffectiveness at higher frequencies, it has been recommended that stranded cables not be used at frequencies over 1200 Hz (5-7). However, in many situations, large cables are required to safely carry currents produced by power faults and lightning discharges; in addition, solid wires larger than approximately 0.6 cm (0.25 in.) may be difficult to obtain.

#### 5.2.4.2 Rectangular Conductors.

At frequencies high enough to make the skin effect noticeable, the resistance ratio of a flat rectangular conductor will be lower than that of a solid round wire with the same cross-sectional area if the width-to-thickness ratio exceeds approximately 2:1. The resistance ratios for several sizes of nonmagnetic ( $\mu_r = 1$ ) rectangular conductors are plotted in Figure 5-6.

The self-inductance at lower frequencies of a rectangular conductor is (5-6)

$$L_{LF} = 0.00508 \ell \left( 2.303 \log \frac{2\ell}{b+c} + 0.5 + .2235 \frac{b+c}{\ell} \right) \mu H, \quad (5-13)$$

where  $\ell$  is the length,  $b$  is the width and  $c$  is the thickness, and all the dimensions are in inches. For the dimensions in cm, Equation 5-13 is

$$L_{LF} = 0.002 \ell \left( 2.303 \log \frac{2\ell}{b+c} + 0.5 + 0.2235 \frac{b+c}{\ell} \right) \mu H. \quad (5-14)$$

If  $\ell$  is larger than 50 ( $b+c$ ), the last term in each equation may be neglected.

The sharp edges on rectangular conductors tend to radiate energy into space and a flat conductor may become an efficient antenna. To reduce the efficiency of the antenna and minimize this radiation, the edges of the rectangular conductor can be rounded to form an elliptical shape.

#### 5.2.4.3 Tubular Conductors.

Tubular conductors provide the best compromise between factors such as availability, cost weight, cross-sectional area, skin effect, resistance ratio and inductance. By using the actual cross-sectional area of the conductive material, the dc resistance of tubular conductors can be determined from Equation 5-1; it is given for three different diameter copper tubes in Figure 5-7.



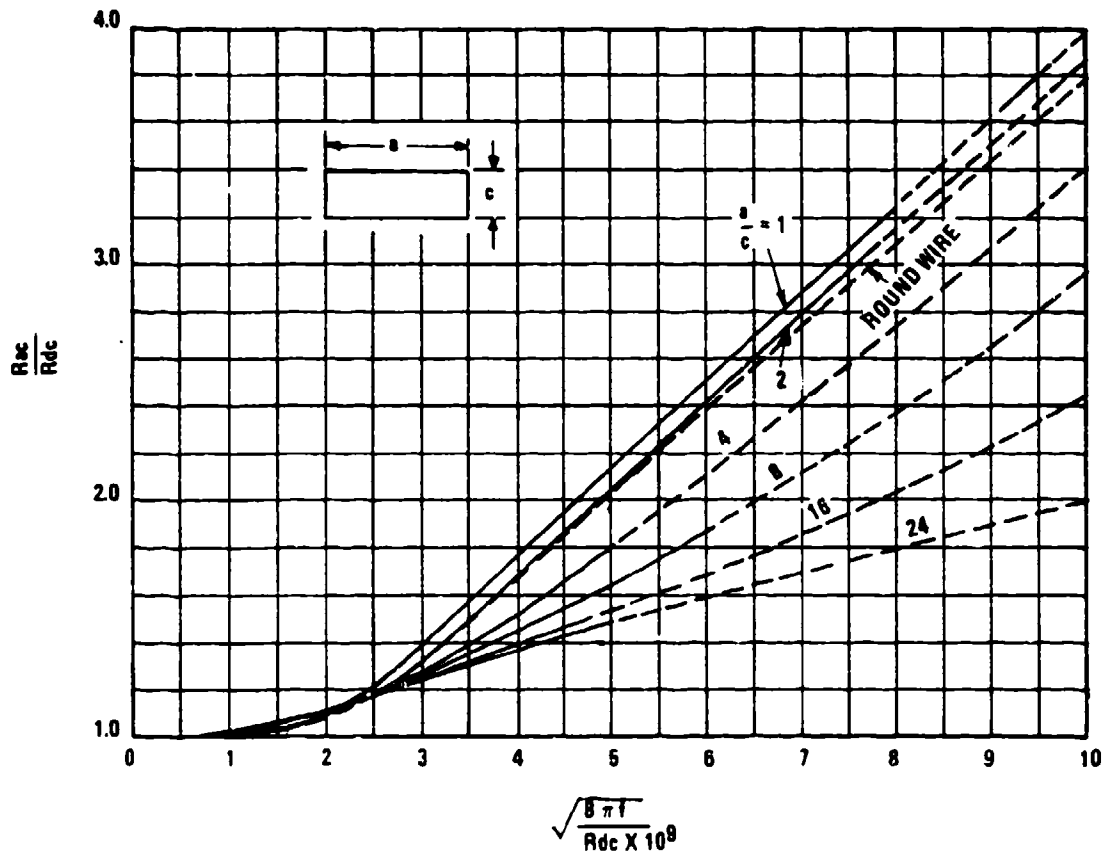


Figure 5-6. Resistance Ratio of Rectangular Conductors (5-3)

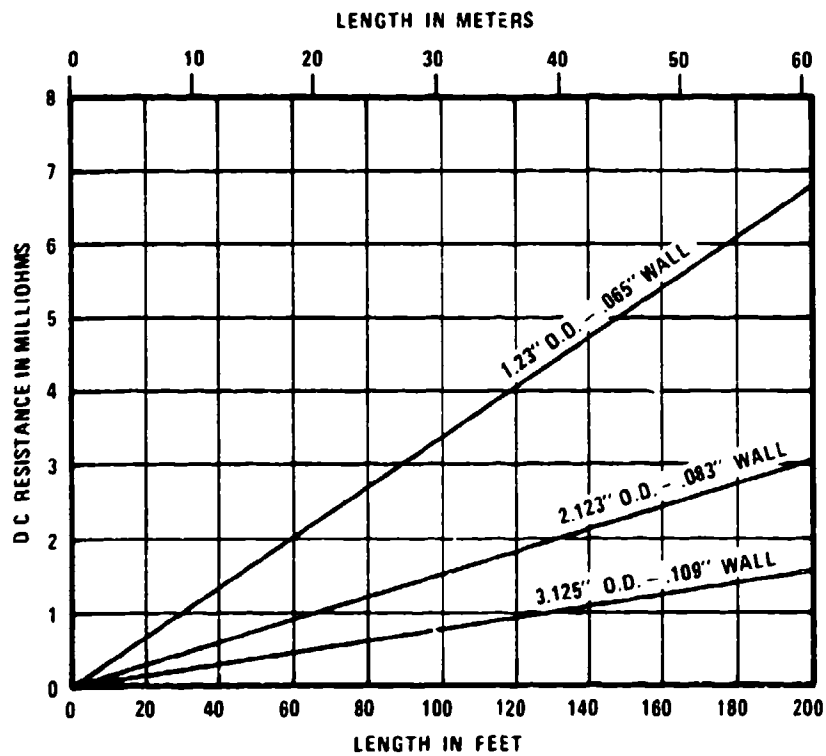


Figure 5-7. Resistance versus Length for Various Sizes of Copper Tubing (5-7)

Both the dc and the ac resistances of a tubular conductor are greater than those of a solid conductor with the same outside diameter. However, the ac resistance does not increase as much as the dc resistance, and, therefore, the resistance ratio of a tubular conductor is always less than that of a solid conductor. The ac resistance for four sizes of copper tubing is given in Figure 5-8, and the resistance ratio for isolated nonmagnetic tubular conductors of various sizes is given in Figure 5-9. For a given length of conductor, the ac resistance per unit weight (i.e., per given amount of copper) is less at high frequencies for tubular conductors than for any other shape.

The self-inductance of a conductor is reduced by the absence of a conductive medium in the center (5-7). Therefore, the self-inductance of a tubular conductor will be less than that of a solid conductor with the same diameter. The self-inductance of three representative sizes of copper tubes is given in Figure 5-10.

**5.2.4.4 Structural Steel Members.** The steel I-beam in the structural framework of a building is another conductor that is frequently used as a ground conductor. The resistivity of steel is approximately ten times that of copper; however, the skin depth of steel is greater than 3 times that of copper. This increased skin depth in steel increases the conducting area for high frequency currents. For example, in comparing a 0.3 meter (12-inch) I-beam with a 4/0 AWG copper cable, the perimeter of the I-beam is about 30 times as great and with a factor of 3 increase in the skin depth, the conducting area for high frequency currents in the steel I-beam is close to 90 times larger. This advantage is offset somewhat by the fact that the current tends to flow in the edges of the I-beam and by the surface roughness. The ac resistance will be increased by a factor of 4 because of this surface roughness and current distribution. Even so, the ac resistance of a 4/0 AWG copper cable is 4.25 times as great as that of a 0.3 meter (12-inch) I-beam. In addition, the building framework usually offers many paths in parallel, thus lowering both the ac resistance and the inductance between any two points (5-8).

**5.3 SIGNAL REFERENCE SUBSYSTEM NETWORK CONFIGURATIONS.** Within a piece of equipment the signal reference subsystem may be a sheet of metal which serves as a signal reference plane for some or all of the circuits in that equipment. Between equipments, where units are distributed throughout the facility, the signal ground network usually consists of a number of interconnected wires, bars or a grid which serves an equipotential plane. Whether serving a collection of circuits within an equipment or serving several equipments within a facility, the signal reference subsystem will be a floating ground, a single-point ground, or a multiple-point ground known as a multipoint or equipotential plane. Of the aforementioned signal reference subsystems, the equipotential plane is the optimum ground for communications-electronics facilities. For existing facilities where the presence of equipment prohibit the installation of an equipotential plane beneath, on, or in the floor, the plane may be installed overhead and the equipment connected to it. It is desirable, but not mandatory, to retrofit existing C-E facilities with equipotential planes.

#### 5.3.1 Floating Ground.

A floating ground is illustrated in Figure 5-11. In a facility, this type of signal ground system is electrically isolated from the building ground and other conductive objects. Hence, noise currents present in the building's ground system will not be conductively coupled to the signal circuits. The floating ground system concept is also employed in equipment design to isolate the signal returns from the equipment cabinets and thus prevent noise currents in the cabinets from coupling directly to the signal circuits.

The effectiveness of floating ground systems depends on their true isolation from other nearby conductors, i.e., to be effective, floating ground systems must really float. In large facilities, it is often difficult to achieve a completely floating system, and even if complete isolation is achieved it is difficult to maintain such a system (5-9).

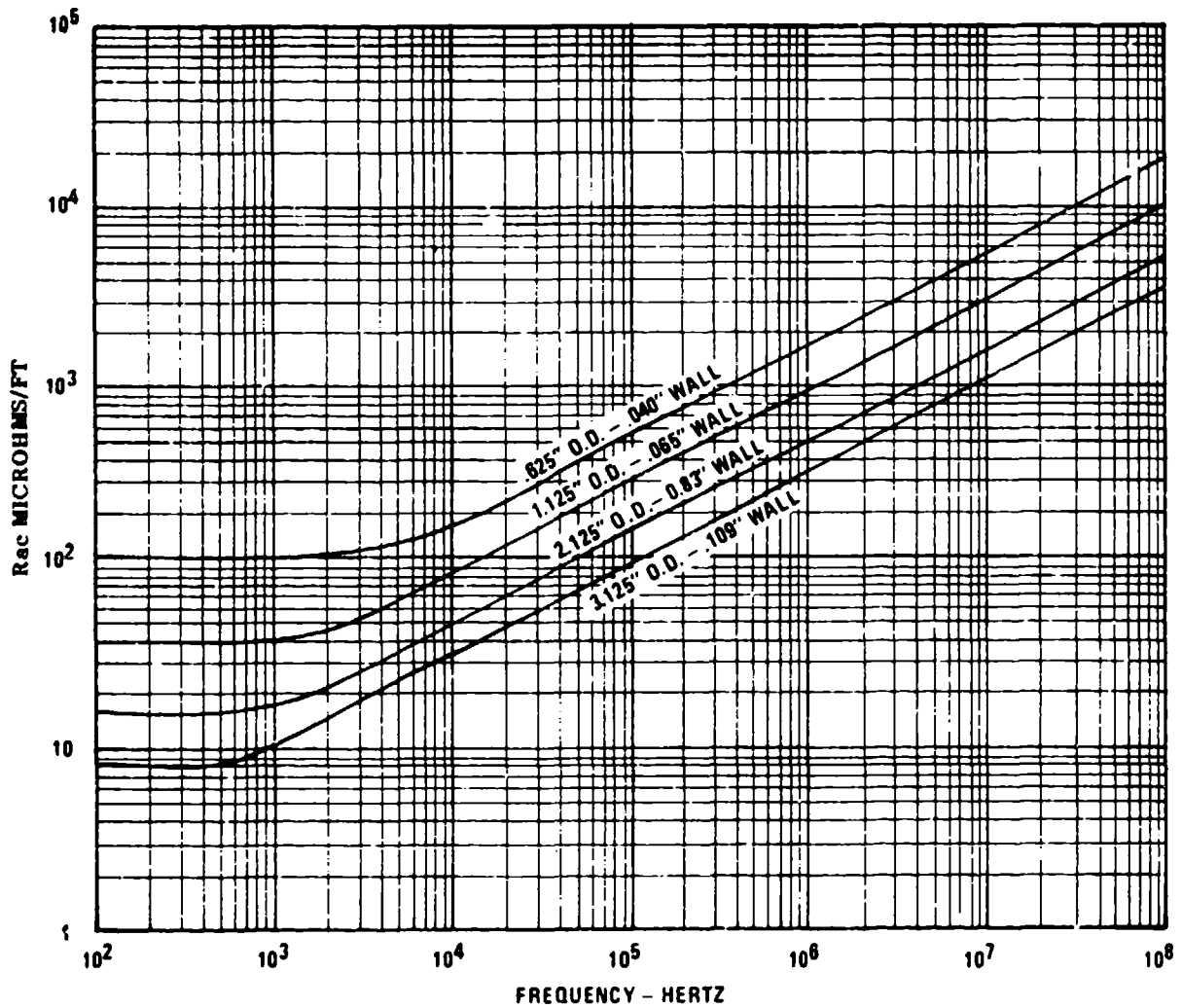


Figure 5-8. AC Resistance versus Frequency for Copper Tubing (5-7)

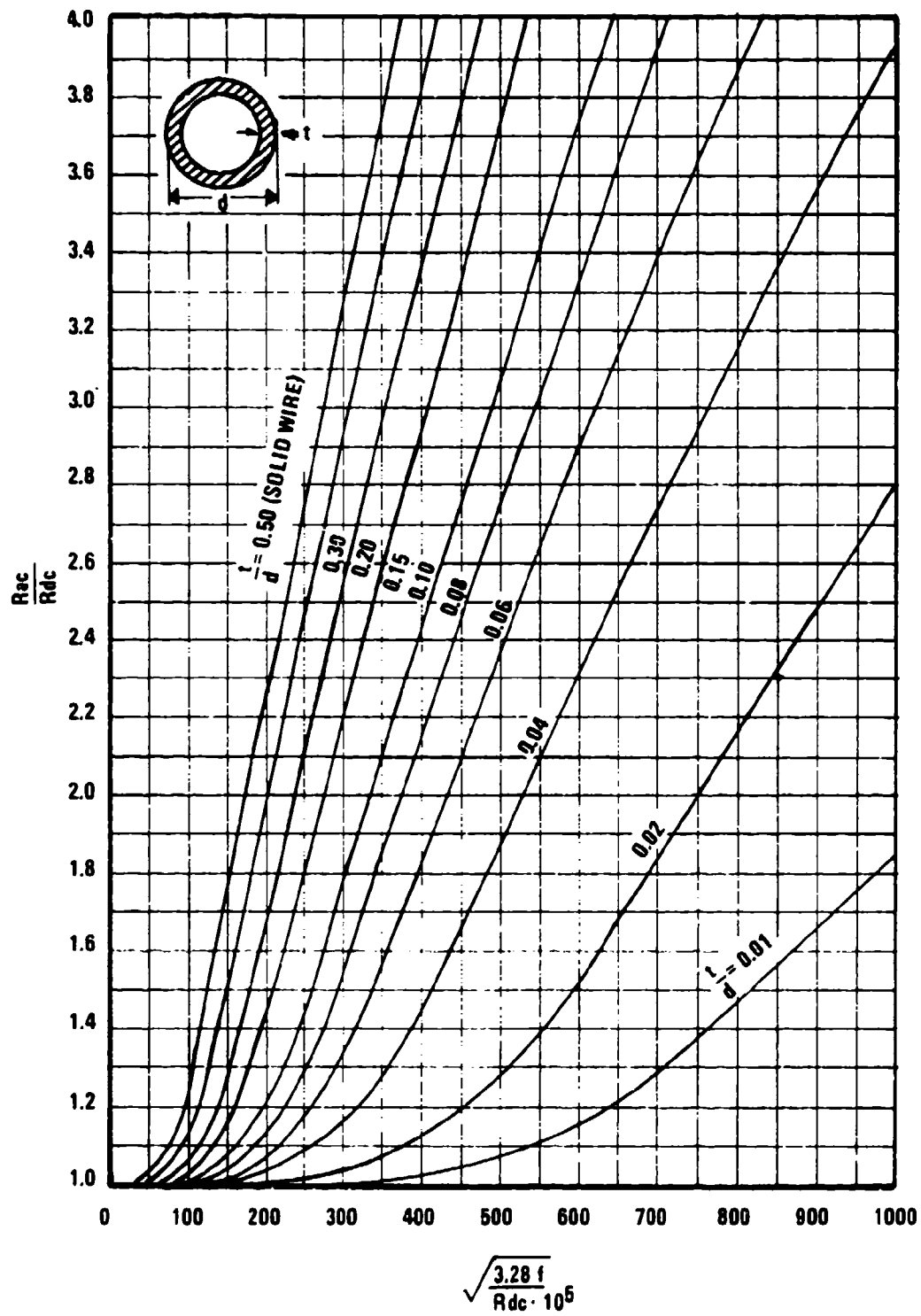


Figure 5-9. Resistance Ratio of Nonmagnetic Tubular Conductors (5-3)

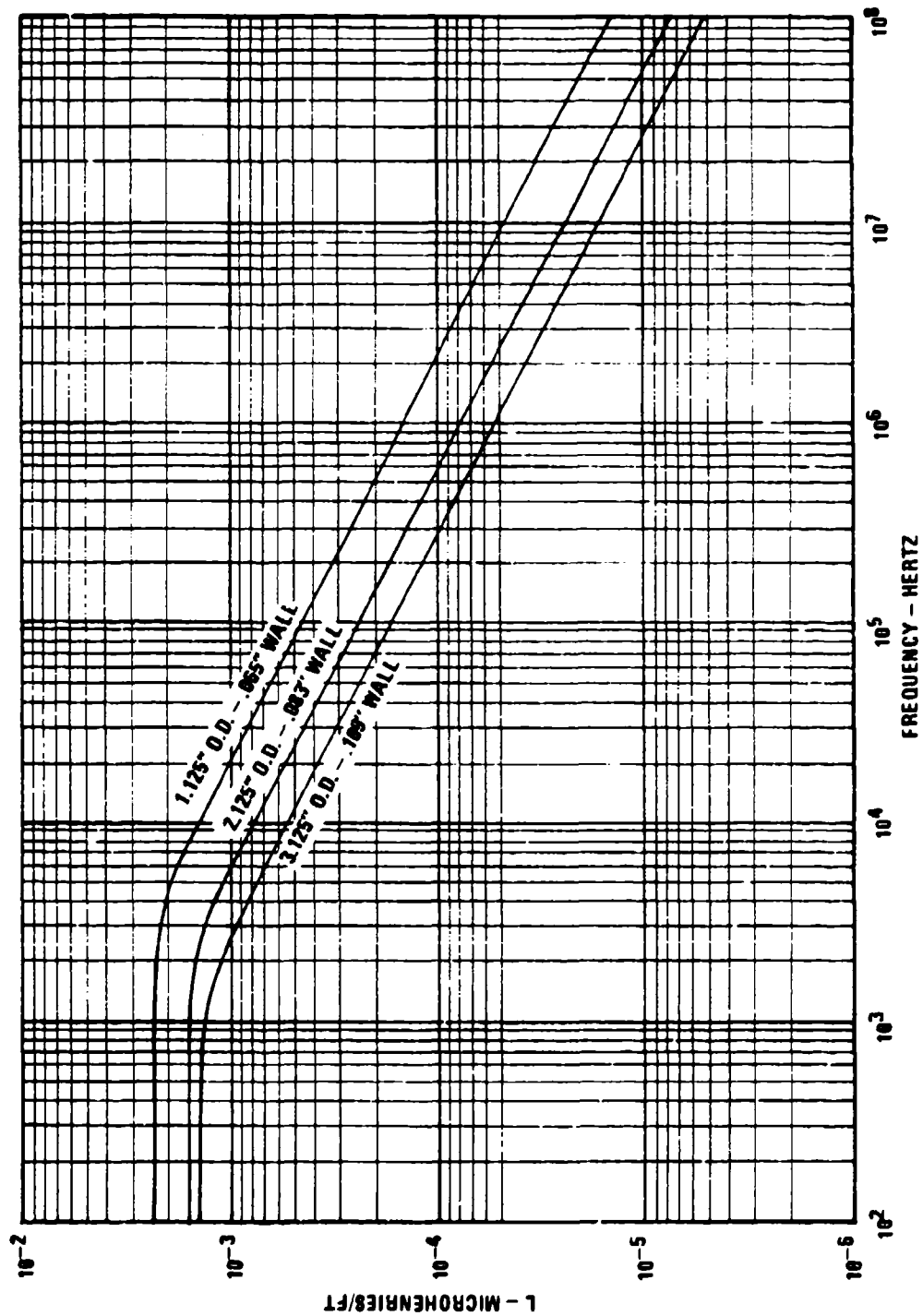


Figure 5-10. Inductance versus Frequency for Various Sizes of Copper Tubing (5-7)

In addition, a floating ground system suffers from other limitations. For example, static charge buildup on the isolated signal circuits is likely and may present a shock and a spark hazard. In particular, if the floated system is located near high voltage power lines, static buildup is very likely. Further, in most modern electronic facilities, all external sources of energy such as commercial power sources are referenced to earth grounds. Thus, a danger with the floating system is that power faults to the signal system would cause the entire system to rise to hazardous voltage levels relative to other conductive objects in the facility. Another danger is the threat of flashover between the structure or cabinet and the signal system in the event of a lightning stroke to the facility. Not being conductively coupled together, the structure could be elevated to a voltage high enough relative to the signal ground to cause insulation breakdown and arcing. This system generally is not recommended for C-E facilities.

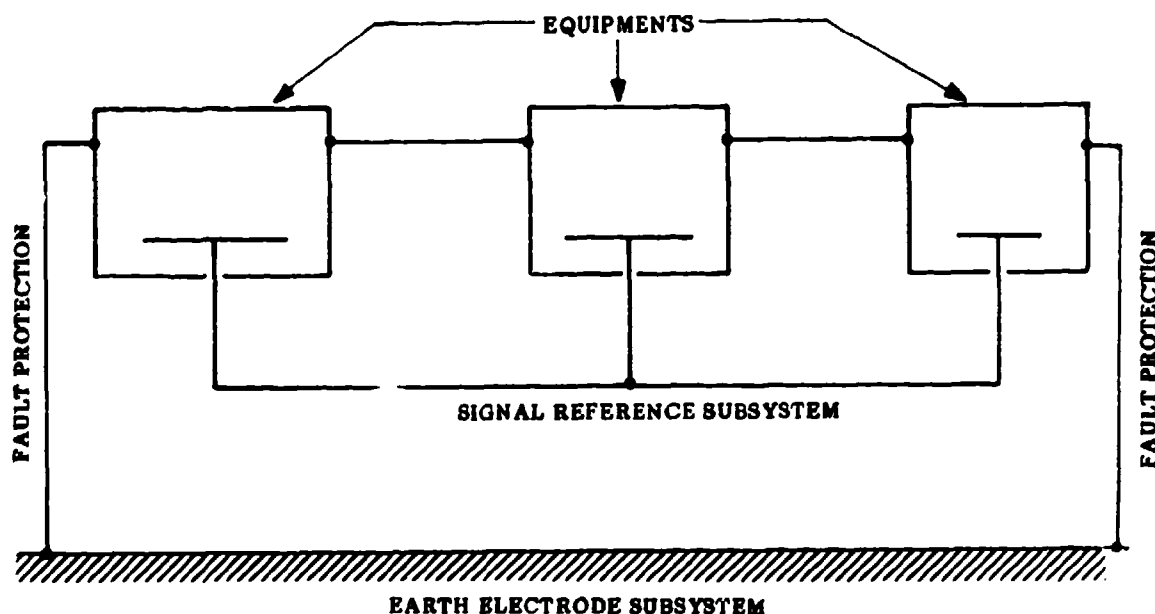


Figure 5-11. Floating Signal Ground

### 5.3.2 Single-Point Ground. (For lower frequencies, 0-30 kHz up to 300 kHz)\*

A second configuration for the signal ground network is the single-point approach illustrated in Figure 5-12. With this configuration, the signal circuits are referenced to a single point, and this single point is then connected to the facility ground. The ideal single-point signal ground network is one in which separate ground conductors extend from one point on the facility ground to the return side of each of the numerous circuits

\* Refer to 5.4.3 for definition of frequency limits.

located throughout a facility. This type of ground network requires an extremely large number of conductors and is not generally economically feasible. In lieu of the ideal, various degrees of approximation to single-point grounding are employed.

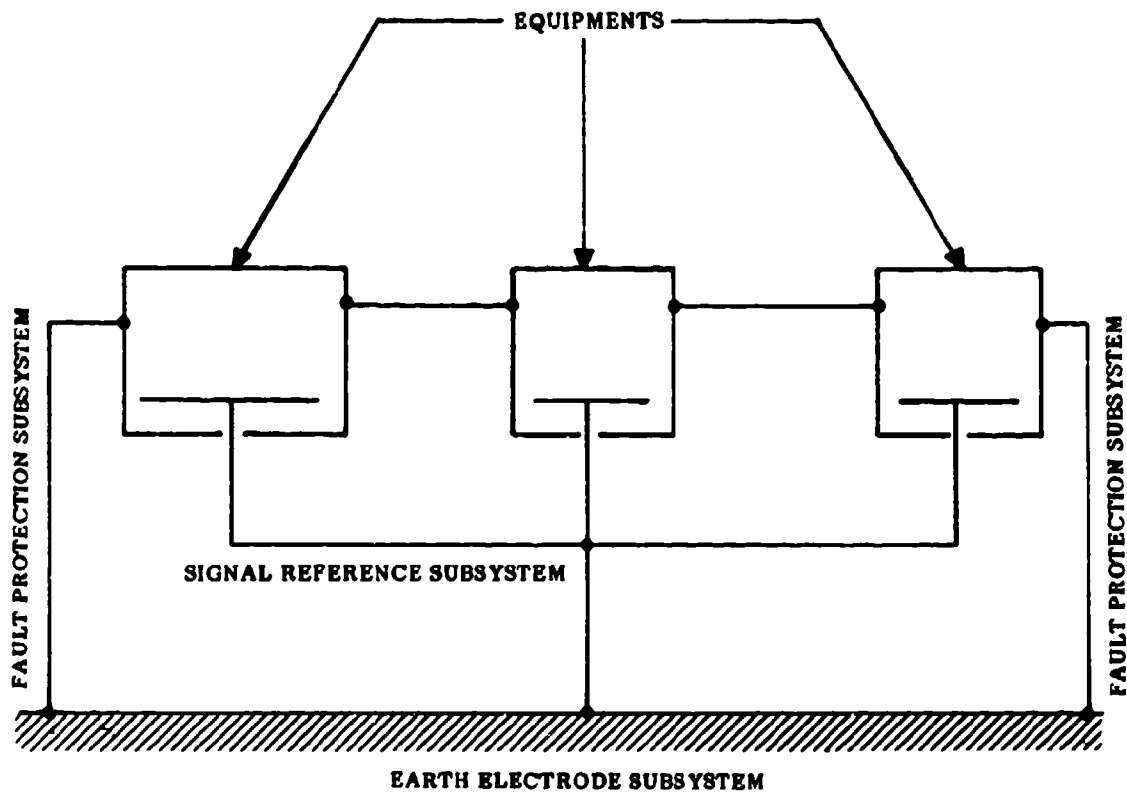


Figure 5-12. Single-Point Signal Ground (For Lower Frequencies)

The configuration illustrated by Figure 5-13 closely approximates an ideal single-point ground. It uses individual ground buses extending from an earth electrode subsystem to each separate electronic system. In each system, the various electronic subsystems are individually connected at only one point to this ground bus. Another frequently used approximation to the ideal is illustrated in Figure 5-14. Here the ground bus network assumes the form of a tree. Within each system, each subsystem is single-point grounded. Each of the system ground points is then connected to a tree ground bus with a single insulated conductor (usually yellow).

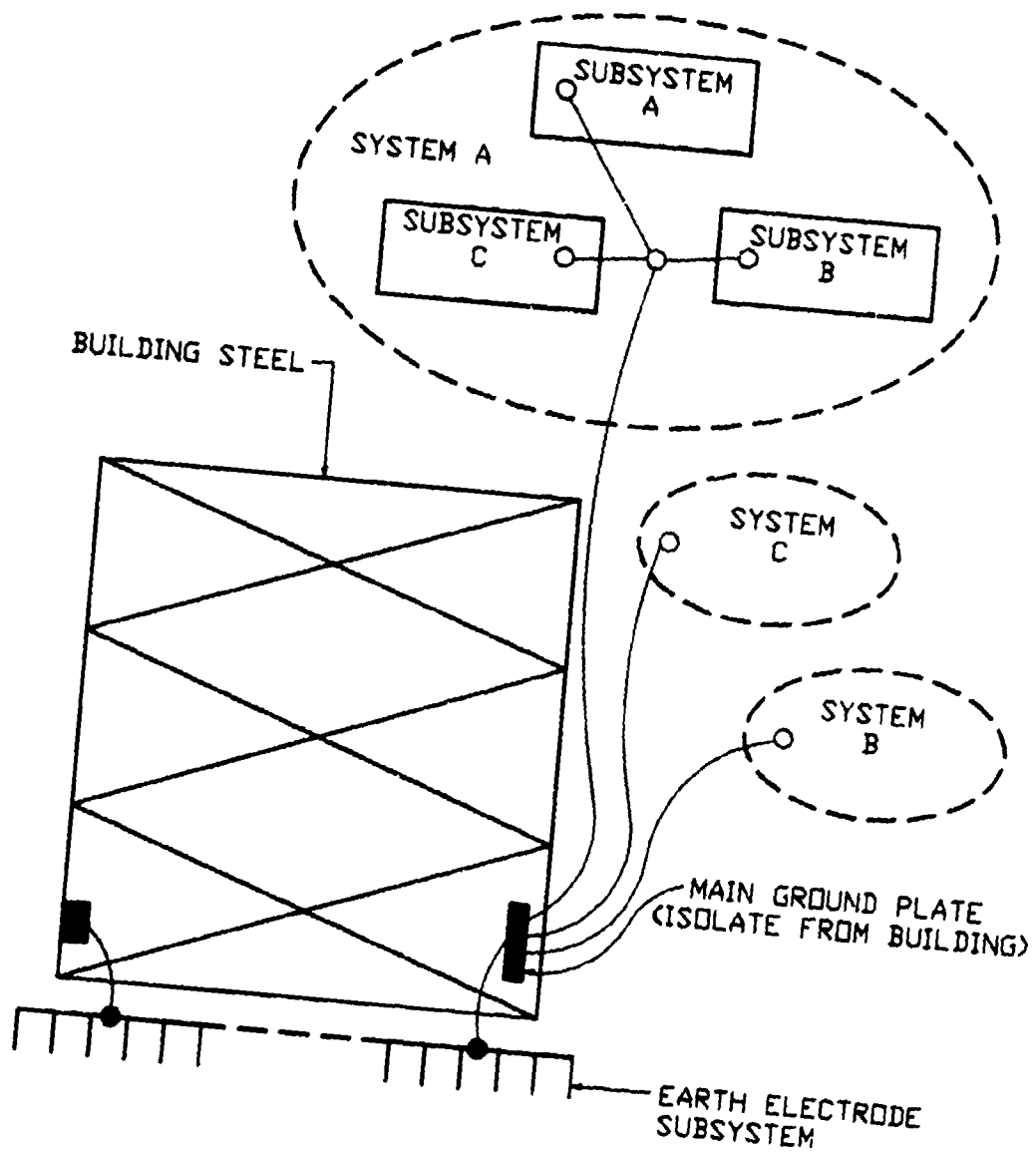


Figure 5-13. Single-Point Ground Bus System Using Separate Risers (Lower Frequency)



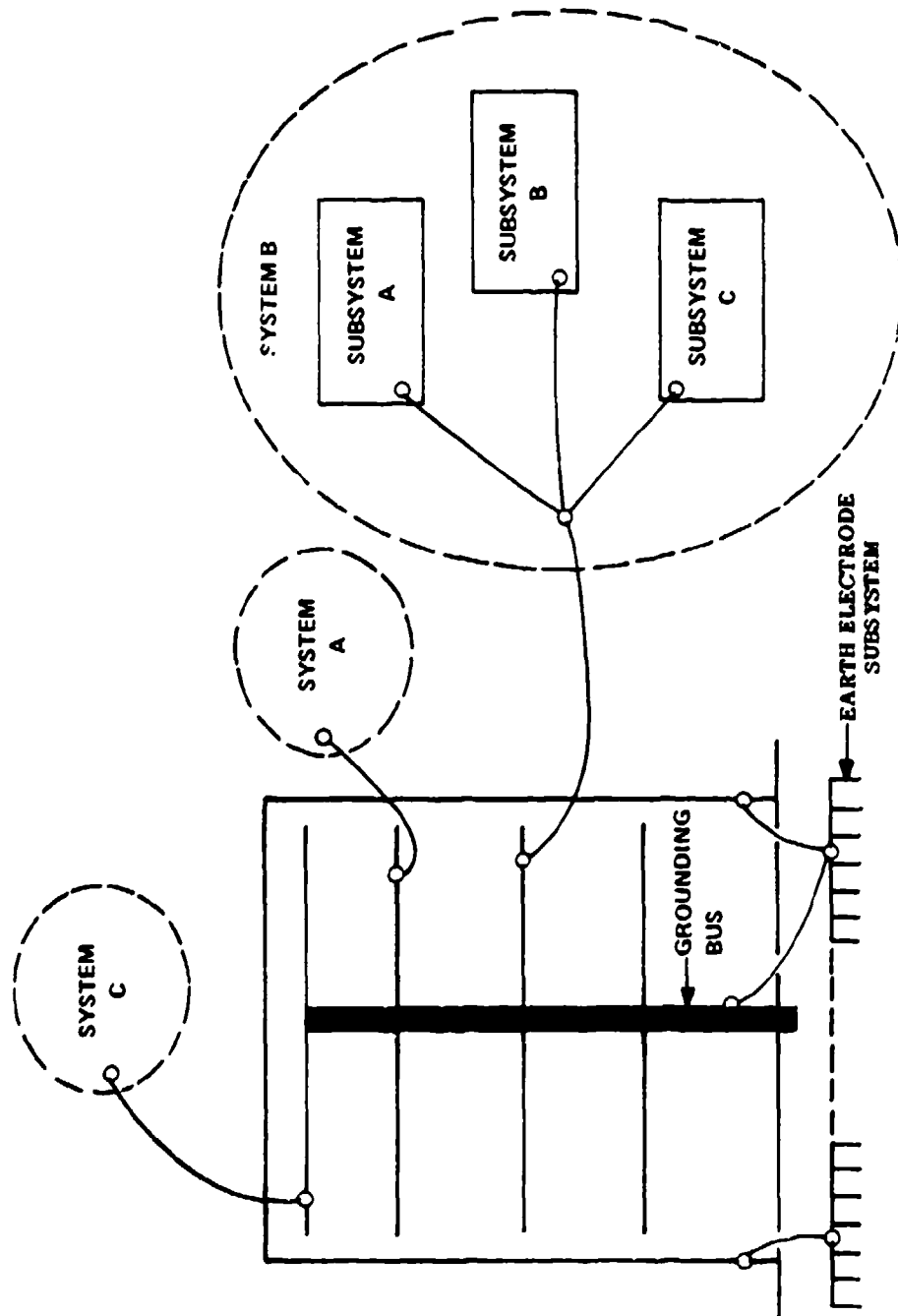


Figure 5-14. Single-Point Ground Bus System Using a Common Bus

The single-point ground accomplishes each of the three functions of signal circuit grounding mentioned at the beginning of this chapter. That is, a signal reference plane is established in each unit or piece of equipment and these individual reference planes are connected together and to the earth electrode subsystem. An important advantage of the single-point configuration is that it helps control conductively-coupled interference. As illustrated in Figure 5-15, closed paths for noise currents in the signal ground network are avoided, and the interference voltage,  $V_N$ , in the facility ground system is not conductively coupled into the signal circuits via the signal ground network. Therefore, the single-point signal ground network minimizes the effects of lower frequency noise currents which may be flowing in the facility ground.

Single-point grounds, however, also become transmission lines at higher frequencies with earth being the other side of the line. In addition, every piece of equipment bonded to this transmission line will act as a tuned stub. In the presence of digital signals (square waves) the tuned circuits will ring at the specific frequencies to which they are resonant. Since single-point grounds behave as transmission lines at rf frequencies, they will have different impedances as a function of frequency, i.e., they may appear as inductors, capacitors, tuned circuits, insulators or pure resistance, and therefore become extremely poor grounds. In a large installation, another major disadvantage of the single-point ground configuration is the requirement for long conductors. The long conductors ( $1/8 \lambda$  at the highest frequency of concern) prevent the realization of a satisfactory reference for

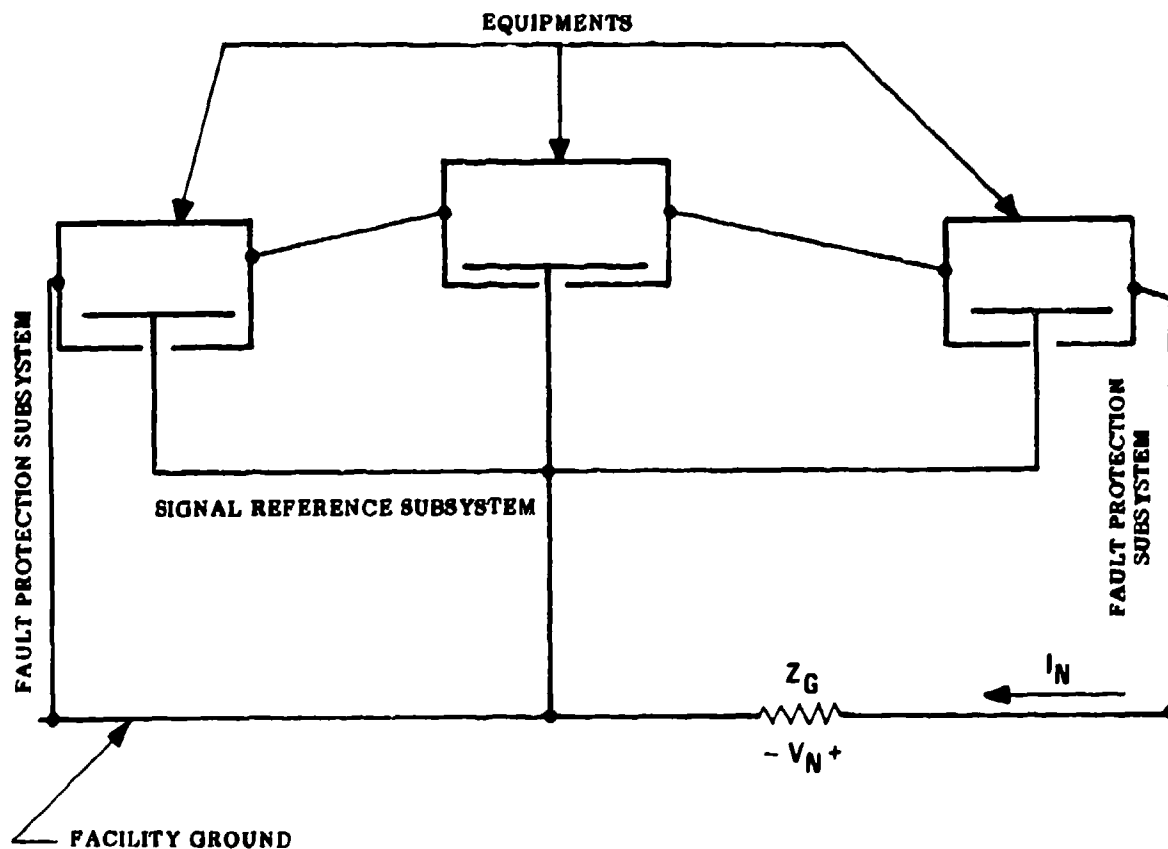


Figure 5-15. Use of Single-Point Ground Configuration to Minimize Effect of Facility Ground Currents

higher frequencies because of large self-impedances. Further, because of stray capacitance between conductors, single-point grounding essentially ceases to exist as the signal frequency is increased (5-10). Because of the aforementioned reasons, single-point grounds are not recommended for use in communications electronics facilities.

### 5.3.3 Multipoint Ground. (For higher frequencies, 30-300 kHz and above)

The multipoint ground illustrated in Figure 5-16 is the third configuration used for signal ground networks. The multipoint ground utilizes many conductive paths from the earth electrode subsystem to various electronic systems or subsystems within the facility. Within each subsystem, circuits and networks are multiply connected to this ground network. Thus, in a facility, numerous parallel paths exist between any two points in the ground network as shown in Figure 5-17.

Multipoint grounding frequently simplifies circuit construction inside complex equipments; it is the only realistic method for the grounding of higher frequency signal circuits. This method of grounding permits equipments employing coaxial cables to be more easily interfaced since the outer conductor of the coaxial cable does not have to be floated relative to the equipment cabinet or enclosure. The multipoint grounding has the disadvantage of exhibiting transmission line characteristics at rf frequencies. To be effective, a multipoint ground system requires an equipotential ground plane whenever the conductors exceed  $1/8 \lambda$  at the highest frequency of concern (5-11).

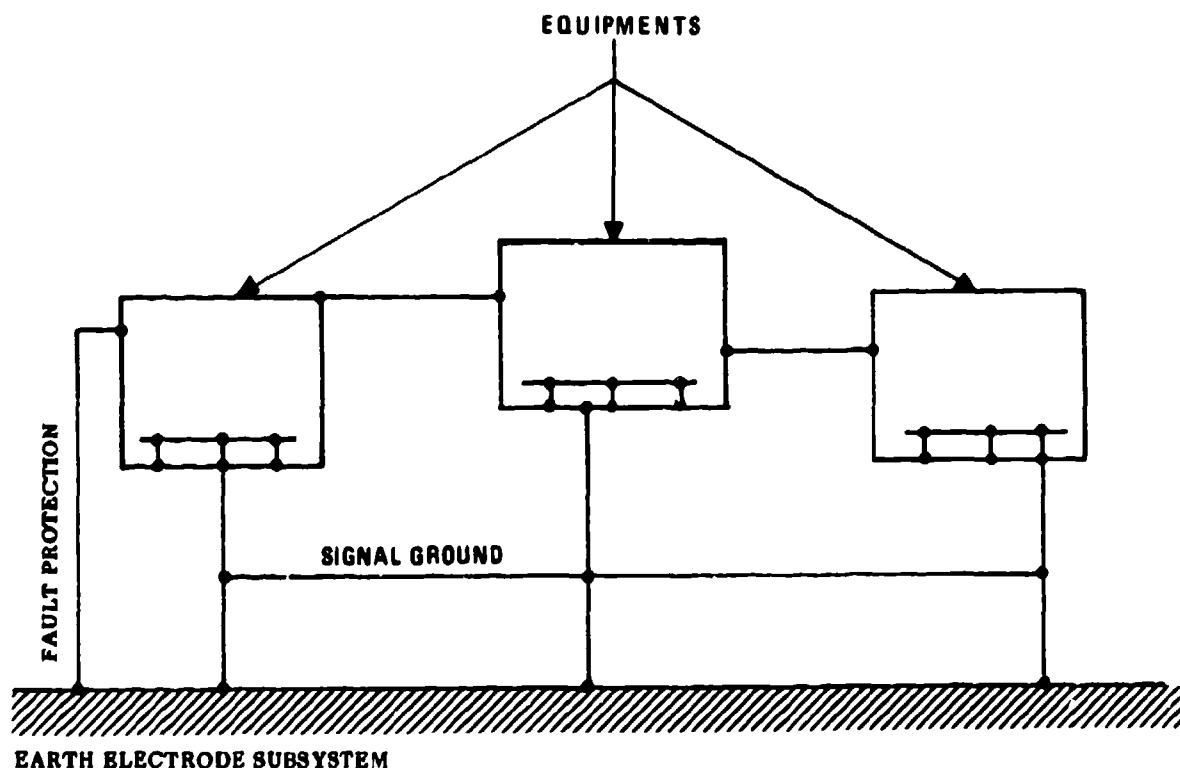


Figure 5-18. Multipoint Ground Configuration

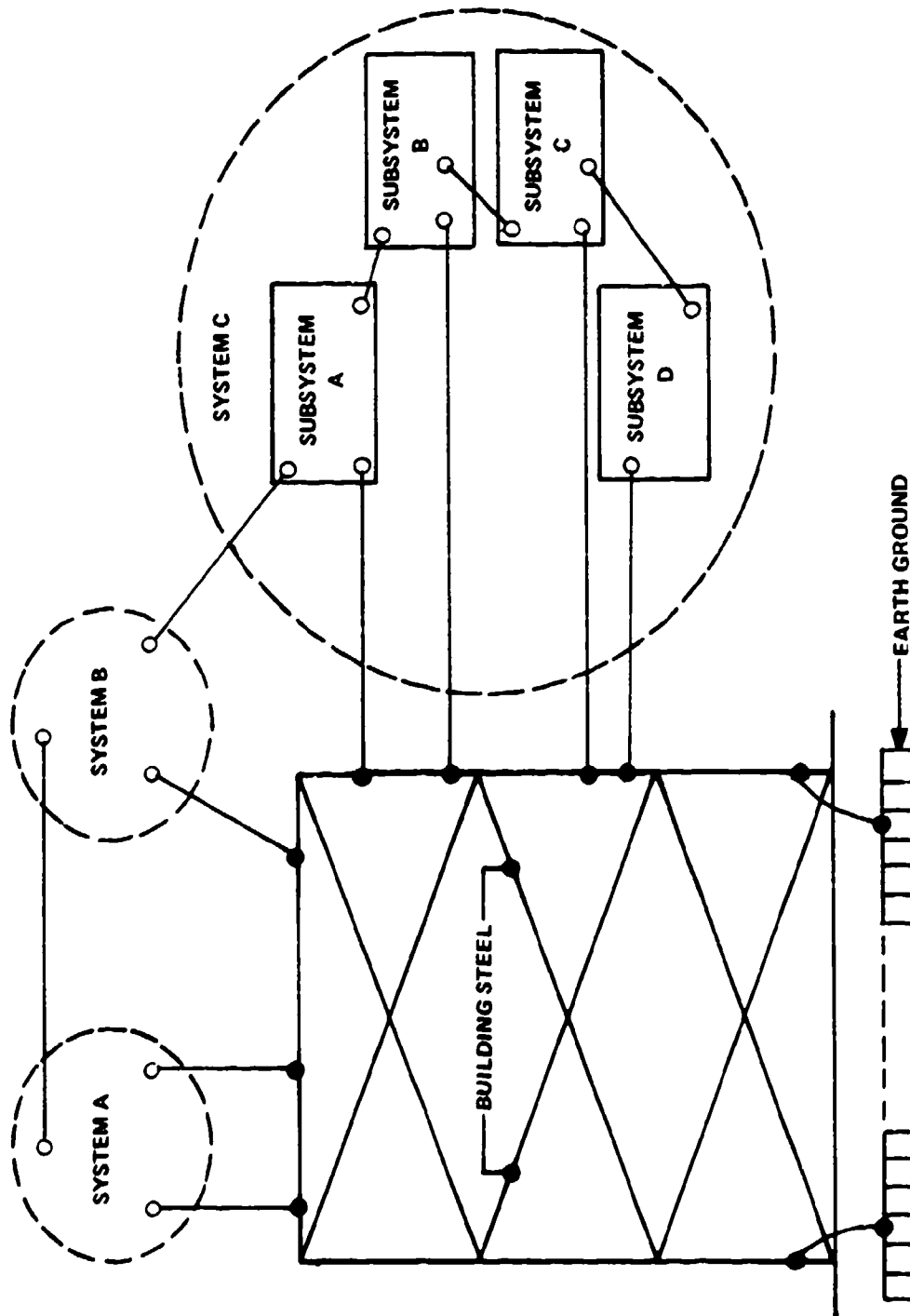


Figure 5-17. Use of Structural Steel in Multiple-Point Grounding

Care must also be taken to ensure sixty hertz power currents and other high amplitude lower frequency currents flowing through the facility ground system do not conductively couple into signal circuits and create intolerable interference in susceptible lower frequency circuits.

#### 5.3.3.1 Equipotential Plane.

The importance of equipotential ground planes cannot be overemphasized for proper equipment operation, as well as for EMI and noise/static suppression. An equipotential ground plane implies a mass, or masses of conducting material which, when bonded together, offers a negligible impedance to current flow. Connections between conducting materials which offer a significant impedance to current flow, can place an equipotential plane at a high potential with respect to earth. High impedance interconnections between metallic members subject to large amounts of current due to power system faults can be extremely hazardous to personnel and equipment. The RFI effect of an equipotential plane or system must however be carefully considered, and it is important to understand that grounding may not, in and of itself, reduce all types of RFI. On the contrary, grounding a system may in some instances increase interference by providing conductive coupling paths or radiative or inductive loops.

Many of the deficiencies of the wire distribution system can be overcome by embedding a large conducting medium, in the floor under the equipments to be grounded. For existing facilities this system may be installed above the equipment to be grounded. A large conducting surface presents a much lower characteristic impedance than that of wire because the characteristic impedance ( $Z_0$ ) is a function of  $L/C$ , hence as capacity to earth increases,  $Z_0$  decreases. The capacity of a metallic sheet or grid to earth is much higher than that of wire. If the size of the sheet is increased and allowed to encompass more area, the capacitance increases. Also, the unit length inductance decreases with width, which further decreases  $Z_0$ . If the dimensions of a metallic sheet increase extensively (as in the case of conducting floor), the characteristic impedance approaches a very low value. In this case, the characteristic impedance would be quite low throughout a large portion of the spectrum. This, in turn, would establish an equipotential reference plane for all equipments bonded to it.

Although it is not necessary from a functional point of view, terminating the surface to an earth connection presents the following advantages:

- a. Personnel safety is not dependent on long cable runs for protection against power faults.
- b. Low impedance is provided for power and radio frequencies.

Grounding buses in a communication facility where higher frequencies are present, act as lossy transmission lines and therefore must be treated as such. Due to this phenomena single-point grounds and multipoint grounds employing ground buses are high impedance grounds at higher frequencies. To be effective at the higher frequencies, the multipoint ground system requires the existence of an equipotential ground plane. Equipotential Planes are sometimes considered to exist in a building with a metal floor or ceiling grid electrically bonded together, or in a building with the ground grid embedded in a concrete floor connected to the structural steel and the facility ground system. Equipment cabinets are then connected to the equipotential plane. Chassis are connected to the equipment cabinets and all components, signal return leads,

etc., are connected to the chassis. The equipotential plane is then terminated to the earth electrode subsystem and to the main structural steel via multiple connections, to assure personnel safety and a low impedance path for all frequencies and signals. It is again emphasized, however, that care must be taken not to create loops which can couple signals from one system to another.

The equipotential plane also offers the following additional advantages;

a. Any "noisy" cable or conductor connected to the receptor, i.e., receivers, modems, etc., through or along such a ground plane will have its field contained between the conductor and the ground plane. The noise field can be "shorted out" by filters and bond straps because the distance between these "transmission line" conductors is very small. Shorting out the noise field has the desirable effect of keeping noise current from flowing over the receptor case and along any antenna input cables.

b. Filters at the interface terminals of equipment can operate more effectively when both terminals of their equivalent "transmission line" are available. As in a, above, a large conducting surface makes it possible to contain the field carried by the offending conductor, in such a way that it can be more easily prevented from traveling further.

c. A large conducting surface may also shield or isolate rooftop antennas from noisy cables below it.

**5.3.3.2 Types of Equipotential Planes.** Conducting materials that can be utilized for equipotential planes are (a) a copper grid embedded in the concrete floor such as a computer floor, (b) a subfloor of aluminum, copper, phosphor bronze screen or sheet metal laid underneath the floor tile or carpet or (c) a ceiling grid above the equipment. Additional data and information on each of these planes can be found in para 1.5.1.1.1 of Vol II.

### 5.3.4 Floating System.

The floating ground system is completely insulated from the building or from any wiring that may be a source of circulating currents. The effectiveness of floating ground systems depends on their true isolation. In large systems, it is difficult to provide required isolation to maintain a good quality floating ground. Insulation breakdown occurs easily because static charges, fault potentials and lightning potentials may accumulate between the floating ground and other accessible grounds, such as external power line neutrals, water pipes, etc. Due to the personnel hazards from the difference of potential between the floating ground and building ground, this system is not recommended.

The preferred grounding method is to have an equipotential plane bonded to the earth electrode subsystem and building structure steel at multiple points with the structural steel also bonded to the earth electrode subsystem. In those facilities which do not have structural steel, multiple copper downloads should be connected from the equipotential plane to the earth electrode subsystem.

## 5.4 SITE APPLICATIONS.

Because of the interference threat that stray power currents present to audio, digital, and control circuits (or others whose operating band extends down to 60 Hz or below), steps must be taken to isolate these large currents from signal return paths. Obviously, one way of lessening the effects of large power currents is to configure the signal ground system so that the signal return path does not share a path common with a power return. This can be accomplished by making sure that the grounding conductor (green wire) of the power system is always run in the same cable, conduit, duct, or raceway with the phase and neutral conductors to the first service disconnect and then bonded to the earth electrode subsystem.

The first step in the development of an interference-free signal reference subsystem for an equipment or a facility is to assure that the ac primary power return lines are interconnected with the safety grounding network at only one point. Isolation of ac power returns from the signal reference subsystem is a major factor toward reducing many noise problems. Additional steps should also be taken to minimize other stray ac currents such as those resulting from power line filters. (One way of reducing these currents is to limit the number of filter capacitors in an installation by using common filtered ac lines wherever possible or by locating the filters as near as possible to the power service entry of the facility.)

To meet the safety requirements while minimizing the effects of power currents flowing with signal currents through a common impedance, a single connection\* between the power distribution neutral and the earth electrode subsystem is necessary. This single connection eliminates conductive loops in which circulating (power) currents can flow to produce interference between elements of the signal reference network.

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\*This connection to the earth electrode subsystem should be made from the first service disconnect. Care should be taken to ensure that the signal reference, fault protection, and lightning protection subsystems are bonded to the earth electrode subsystem at separate ground rod locations.

**5.4.1 Lower Frequency Network (0-30 kHz, and in some cases up to 300 kHz).** The lower frequency grounding network for the facility should conform to the following principles:

a. It should be isolated from other ground networks including structural, safety, lightning and power grounds, etc. The purpose of this isolation is to prevent stray currents (primarily 50/60 Hz power) from developing voltage differentials between points on the ground network.

b. The inter-equipment or facility ground system should not be expected to provide the primary return path for signal currents from the load to the source. Figure 5-18 illustrates a way of discriminating against those extraneous signals which may inductively or capacitively induce currents into the grounding network and develop differential voltages between the source and the load. For example, Figure 5-19 illustrates a practice that is not recommended. If only one source and one load constitute the entire system or if the various source-load pairs within the system are essentially noninterfering in nature of their operation, this grounding arrangement may be acceptable.

c. The lower frequency grounding network must be connected to the earth electrode subsystem at only one point.

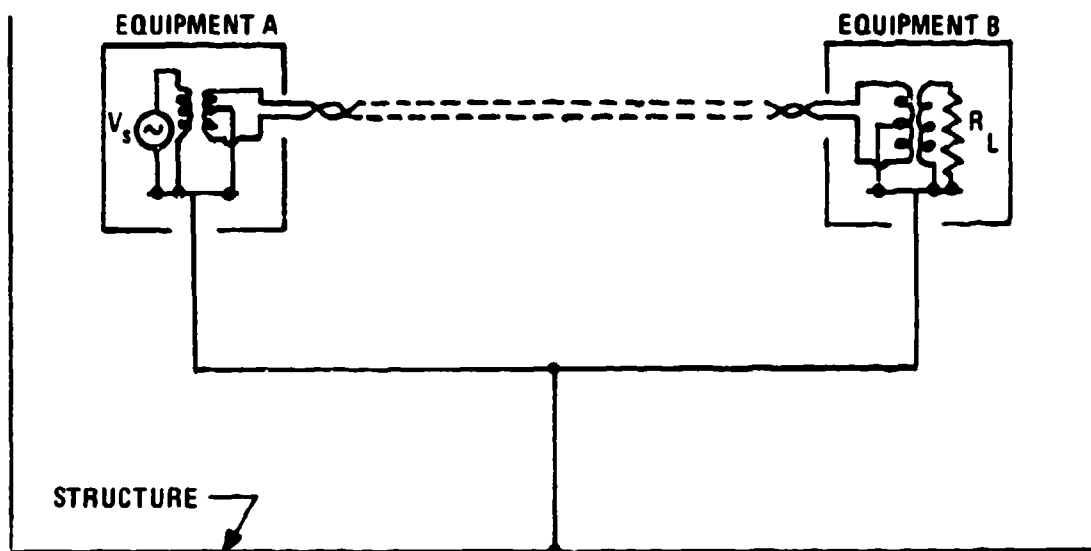


Figure 5-18. Recommended Signal Coupling Practice for Lower Frequency Equipment



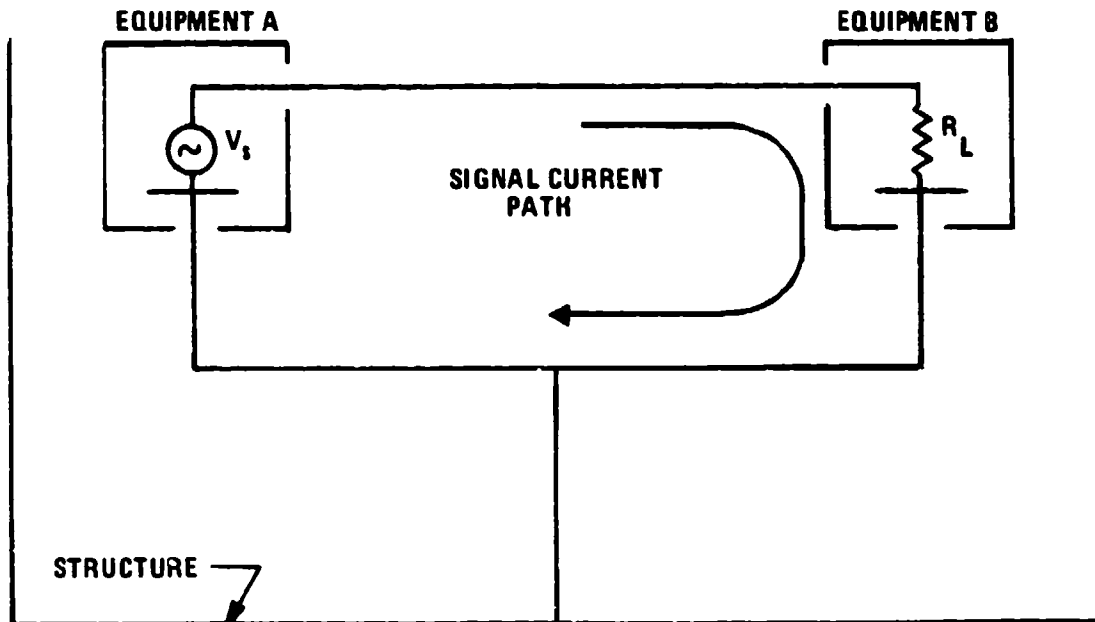


Figure 5-19. Ground Network Used as Signal Return  
(Practice Not Generally Recommended)

d. The network must be configured to minimize conductor path lengths. In facilities where the equipments to be connected to the ground network are widely separated, more than one network should be installed.

e. Finally, the conductors of the network are to be routed in a manner that avoids long runs parallel to primary power conductors, lightning down conductors, or any other conductor likely to be carrying high amplitude currents.

#### 5.4.2 Higher Frequency Network (>300 kHz, and in some cases down to 30 kHz).

The higher frequency (equipotential) network provides an equal potential plane with the minimum impedance between the associated electronic components, racks, frames, etc. This plane shall be used at facilities or areas within facilities where interface frequencies are over 300 kHz and may be used at sites where interface frequencies are as low as 30 kHz. In higher frequency systems, equipment chassis are frequently used as the signal reference. The chassis in turn is usually connected to the equipment case at a large number of points to achieve a low impedance path at the frequencies of interest. See Para. 5.4.3.

The National Electrical Code requires that equipment cases and housings be grounded to protect personnel from hazardous voltages in the event of an electrical fault. Stray currents in the fault protection network can present an interference threat to any signal system whose operating range extends down into the lower frequency range and should be eliminated. Where such problems exist, it is advisable to attempt to reduce the impedance of the reference plane as much as possible. A practical approach is to interconnect equipment enclosures with the equipotential plane, via building structural steel, cable trays, conduit, heating ducts, piping, etc., into the earth electrode subsystem to form as many parallel paths as possible. It should be recognized that because of the inductance and capacitance of the network conductors, such multipoint ground systems offer a low impedance only to the lower frequency noise currents; however, these currents can be the most troublesome in many facilities. Higher frequencies find a much lower impedance to ground through the distributed capacity of the equipotential plane.

#### 5.4.3 Frequency Limits.

The question remaining concerns the frequency below which signals can be considered as lower frequency. Certainly the dividing line between the lower and higher frequency should be high enough to include all audio communications signals. Since digital systems employ frequencies which extend from dc up to several hundred MHz, a decision based on pulsed-signal considerations is more appropriate. To minimize the possibility that the ground bus conductors will form antennas, the lengths should not exceed 0.02 wavelength which is approximately 21 meters (70 feet) at 300 kHz. Since the grounding buses in medium to large sized facilities may extend 21 meters (70 feet), 300 kHz appears to be the maximum frequency for which a single-point grounding system should be used. At frequencies up to 30 kHz, conductor lengths up to 210 meters (700 feet) can be approached without exceeding the 0.02 wavelength criteria.

MIL-STD-188-124A establishes the lower frequency network range from dc to 30 kHz and in some cases (depending on the interface frequency) up to 300 kHz. The higher frequency network range extends above 300 kHz and may in some cases be used at sites where the interface frequencies are as low as 30 kHz. The frequency range from 30 kHz to 300 kHz is a mutual area and may be considered as either higher or lower depending upon the interface frequency.

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## CHAPTER 6

## INTERFERENCE COUPLING AND REDUCTION

6.1 INTRODUCTION.

A large number of diverse equipments are usually present in an electronics complex. The various systems and subsystems making up the complex may be concentrated in a small area such as a single room or they may be distributed over a wide geographical area and be located in several buildings. Whether the distances between individual equipments are large or small, the entire system must function as an integral unit. Each equipment must supply its designated output -- whether it be audio or rf, or analog or digital -- to some terminal point such as an antenna, land line, or another piece of equipment. Both primary and backup power must be supplied. Critical points in the system must be monitored both locally and remotely. To perform all the required tasks and functions, many control, power distribution, and signal transmission networks are necessary.

Within the interconnected complex, many potentially incompatible signals are present. For example, at one extreme are the large power sources (primarily dc and 60 Hz) supplying the various subsystems. At the other extreme, low level dc and very low frequency signals from monitors, indicators and other specialized devices are present. Also in the low frequency range are audio signals used for communications and control functions. In the higher frequency region of the spectrum are the rf signals ranging from hf to microwaves used for communications, surveillance and tracking, and other functions. These signals range in amplitude from the microwatt levels typical at communications receiver inputs to the kilowatt and megawatt levels transmitted by some radar systems. Ranging from audio frequencies into the rf region are the broadband display and communications systems, both analog and digital, which may span from a few hertz to several megahertz in frequency and may range in amplitude from a few millivolts to a few volts. Falling in overlapping frequency ranges, these various signals present within the complex may interact in an undesirable manner to cause interference (generally manifested as annoying "noise").

Interference is any extraneous electrical or electromagnetic disturbance that tends to interfere with the reception of desired signals or that produces undesirable responses in electronic systems. Interference can be produced by both natural and man-made sources either external or internal to the electronic system. The major objective of interference reduction in modern electronic equipments and facilities is to minimize and, if possible, prevent degradation in the performance of the various electronic systems by the interactions of undesired signals, both internal and external.

The correct operation of complex electronic equipments and facilities is inherently dependent on the frequencies and amplitudes of both the signals utilized in the system and the interference signals present in the facility. If the frequency of an undesired signal is within the operating frequency range of the system, errors in the system response may be obtained. The extent of the system response is a function of the amplitude of the undesired signal relative to that of the desired signal. For example, in systems operating with high level signals, undesired signals with amplitudes on the order of volts may be tolerable, while in low level systems a few microvolts may produce intolerable errors in the response of the system. An important element in the control of unwanted interactions between signals is the proper grounding of the system.

An ideal signal system is a simple signal generator-load pair as shown in Figure 6-1. With no extraneous voltages present within the loop, this simple pair is free of interference. Consider, however, what happens when the current return path is non-ideal and sources of noise are present as shown in Figure 6-2. Unless noise voltages  $V_{N1}$  and  $V_{N2}$  are identical, a voltage difference will exist between the low side of the generator (Node 1) and the low side of the load (Node 2). As shown in Figure 6-3, this voltage difference effectively appears in the signal transfer loop in series with the signal generator and produces noise currents in the load. Four ways of combating this noise problem are as follows:

- a. Isolate the source-load pair from the noise sources; i.e., float the system and provide the necessary shielding and filtering to prevent coupling by other means.
- b. Connect the low side of the loop to the reference plane at either Node 1 or Node 2 but not at both.
- c. Reduce the impedance,  $Z_{\text{return}}$ , of the path connecting the two noise sources.
- d. Reduce the magnitudes of  $V_{N1}$  and  $V_{N2}$  through the control of the currents producing them by lowering the impedance through which these currents flow.

Practical electronics circuits typically are a collection of several source-load combinations such as shown in Figure 6-4. These various source-load combinations may be functionally dependent on each other. Hence each individual source-load pair can not operate in isolation; there must be coupling between pairs. For example, one source may be driving several loads; one load may be receiving signals from several sources; or the load for one signal source may serve as the source for another load. At the circuit level, numerous sources and loads are connected in an interrelated fashion and the use of individual return paths for each source-load pair becomes impractical. It is more realistic to establish a common ground or reference plane which serves as the return path for several signals. The control of undesired network responses, particularly in high gain and/or higher frequency circuits, often requires the establishment of a common signal reference to which functional grouping of components, circuits, and networks can be connected. Ideally, this common reference connection offers zero impedance paths to all signals for which it serves as a reference. The several signal currents within the network can then return to their sources without creating unwanted conductive coupling between circuits.

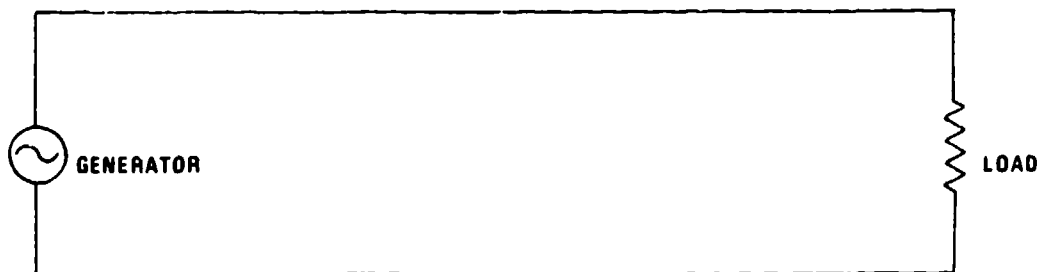


Figure 6-1. Idealized Energy Transfer Loop

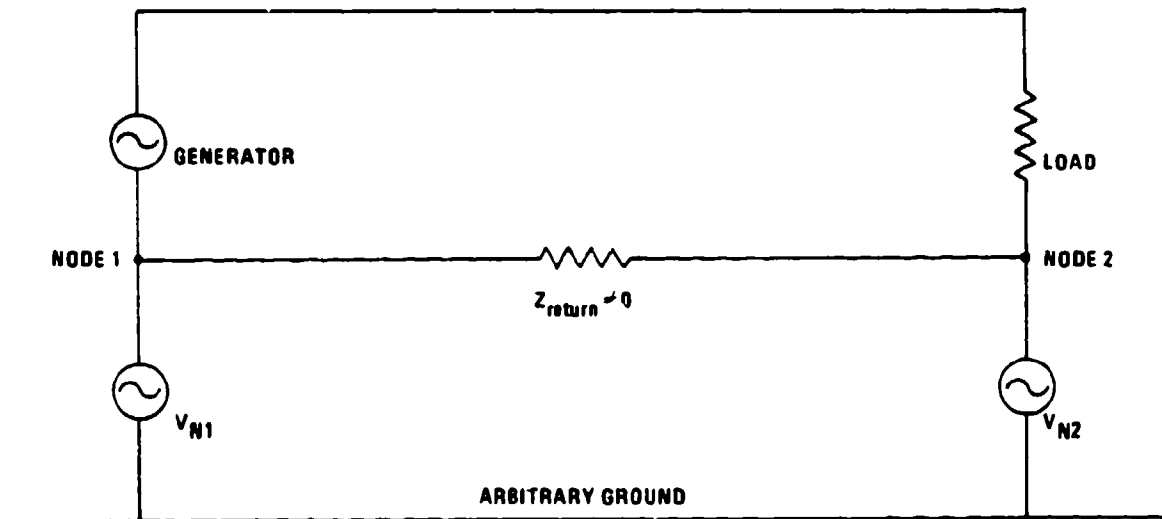


Figure 6-2. Energy Transfer Loop With Noise Sources in Ground System

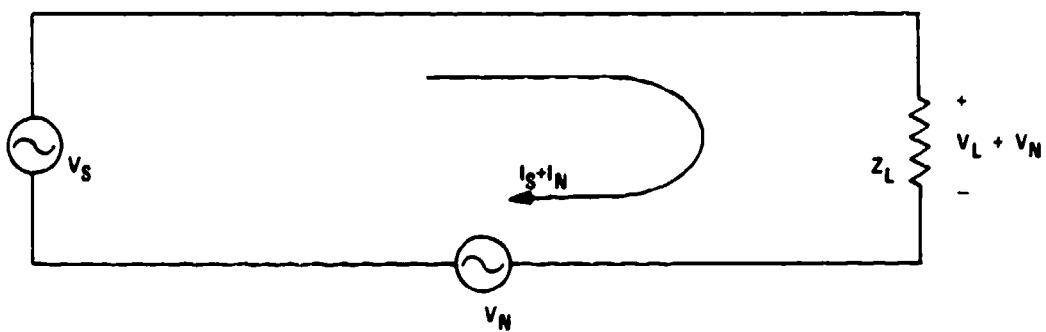
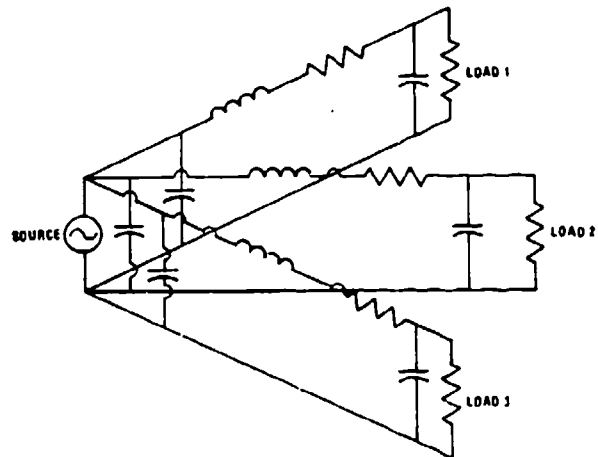
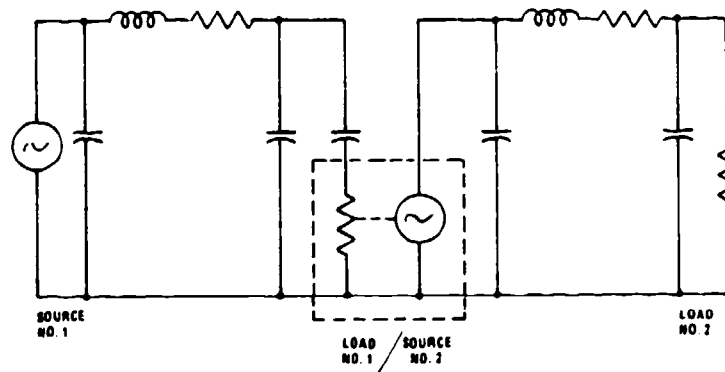


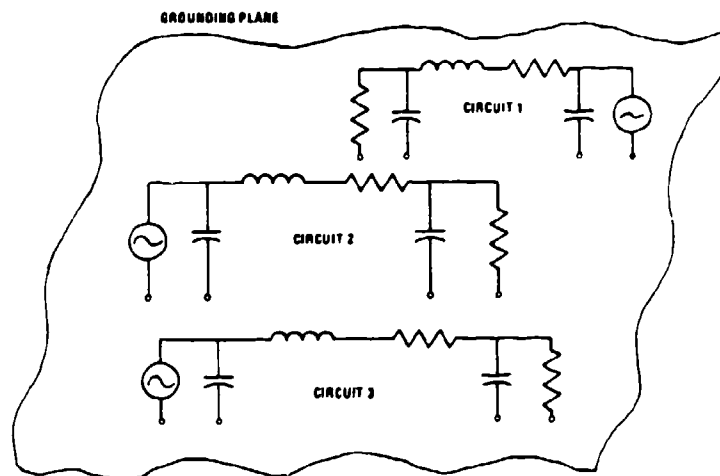
Figure 6-3. Equivalent Circuit of Non-Ideal Energy Transfer Loop



(a) ONE SOURCE DRIVING MULTIPLE LOADS.



(b) CASCADED PAIRS.



(c) SEPARATE PAIRS CONNECTED TO COMMON GROUND PLANE.

Figure 6-4. Practical Combinations of Source-Load Pairs

At the equipment level, the individual signal reference planes for the various networks must be connected together to prevent personnel shock hazards (see Chapters 5 and 9) and to provide as near as possible, the same signal reference for all networks. Thus, the signal reference plane may extend over large distances within a facility. The assumption that this large reference plane provides zero impedance paths is not valid; the series inductance and resistance of the conductors forming the signal reference plane and the shunt capacitance to nearby conductive objects must be considered. Currents flowing in the signal reference plane will develop voltages across this impedance and will produce electric and magnetic fields around the conductors.

## 6.2 COUPLING MECHANISMS.

Coupling is defined as the means by which a magnetic or electric field produced by one circuit induces a voltage or current in another circuit. Interference coupling is the stray or unintentional coupling between circuits which produces an error in the response of one of the circuits. The possible sources of spurious signals and the mechanisms by which this interference is coupled into a susceptible circuit must be understood in order to guard against interference pickup by sensitive signal circuits.

The techniques for reducing pickup depend on the type of interference present. Interference is broadly classified by its coupling means; i.e., as either being conductive or free-space. Conductive coupling occurs when the interfering and the interfered-with circuits are physically connected with a conductor and share a common impedance. Free-space coupling occurs when a circuit or source generates an electromagnetic field that is either radiated and then received by a susceptible circuit or that is inductively or capacitively coupled (near-field) to a susceptible circuit.

### 6.2.1 Conductive Coupling.

Power lines entering a facility provide good conductive coupling paths from interference sources external to the facility. This interference can easily be conducted into a particular unit or piece of equipment via the power lines entering the equipment. Also, interference can conductively couple between various circuits inside the equipment on the common dc power lines. If one dc power supply is utilized with several circuits operating over various signal voltage and frequency ranges, the operation of one circuit may adversely affect the operation of other circuits. For example, if both the preamplifier and the power amplifier sections of an audio amplifier are supplied from a single dc power supply, variations in the relatively large current drawn by the power amplifier can appear as supply voltage variations at the preamplifier. These variations can be large compared to the operating signal levels in the preamplifier; the unwanted variations are amplified along with the desired signals and may produce distortion in the output of the amplifier.

Another set of paths for conductive coupling of interference is offered by the signal lines. In general, signal lines enter each facility and each unit or piece of equipment; i.e., such signal lines are usually necessary for interfacing electronic circuits. Interference can be conductively coupled into facilities, equipment, and circuits as readily by signal lines as by power lines.



The signal reference plane is another potential coupling path for unwanted signals between equipments and/or circuits. Since practical signal reference planes do not exhibit a zero impedance, any currents flowing in a signal reference plane will produce potential differences (voltages) between various points on the reference plane. Interfacing circuits referenced to these various points can experience conductively coupled interference in the manner illustrated by Figure 6-5. The signal current  $I_1$ , flowing in Circuit 1 of Figure 6-5 returns to its source through signal reference impedance  $Z_R$  producing a voltage drop  $E_{N1}$  in the reference plane. The impedance  $Z_R$  is common to Circuit 2, hence  $E_{N1}$  appears in Circuit 2 as a voltage in series with the desired signal voltage source,  $E_{S2}$ . This undesired source produces an interference voltage,  $V_{N2}$ , across the load of Circuit 2; similarly, the desired current,  $I_2$ , in Circuit 2 will produce interference in Circuit 1.

In a facility, the conductive coupling of interference through the signal reference plane of interfaced equipment can occur in a manner similar to that described above for internal circuitry. If Circuit 1 in Figure 6-5 represents two interfaced equipments and if Circuit 2 represents a different pair of interfaced equipments, then a current flowing in either circuit can produce interference in the other circuit as described. Even if the pairs of equipments do not use the signal reference plane as the signal return, the signal reference plane can still be the cause of coupling between equipments. Figure 6-6 illustrates the effect of a stray current,  $I_R$ , flowing in the reference (or ground) plane.  $I_R$  may be the result of the direct coupling of another pair of equipments to the signal reference plane, or it may be the result of free-space coupling to the signal reference plane. In either case,  $I_R$  produces a voltage  $E_N$  in the reference plane impedance,  $Z_R$ . This voltage produces a current in the interconnecting loop which in turn develops a voltage across  $Z_L$  in Equipment B. Thus, it is evident that interference can conductively couple via the signal reference plane to all circuits and equipments connected across the non-zero impedance elements of that reference plane.

#### 6.2.2 Free-Space Coupling.

Free-space coupling is the transfer of electromagnetic energy between two or more circuits not directly interconnected with a conductor. Depending on the distance between the circuits, the coupling is usually defined as either near-field or far-field. Near-field coupling can be subdivided into inductive and capacitive coupling, according to the nature of the electromagnetic field. In inductive coupling, a magnetic field linking the susceptible device or circuit is set up by the interference source or circuit. Capacitive coupling is produced by an electric field between the interference source and the susceptible circuit.

Radiation of energy by electromagnetic waves is the principle coupling mechanism in far-field coupling. The term "radiated coupling" is sometimes used to describe both near-field (inductive and capacitive) coupling and far-field coupling. However, radiated coupling is generally accepted to mean the transfer of energy from a source to a susceptible circuit by means of electromagnetic waves propagating through space according to the laws of wave propagation.

##### 6.2.2.1 Near-Field Coupling.

When two or more wires or other conductors are located near each other, currents and voltages on one wire will be inductively and capacitively coupled to the other wires. The wire acting as the interference source for this near-field coupling may be any conductor such as a high level signal line, an ac power line, a control line, or even a lightning down conductor. The currents or voltages induced into the other wires can further be conductively coupled into susceptible circuits.

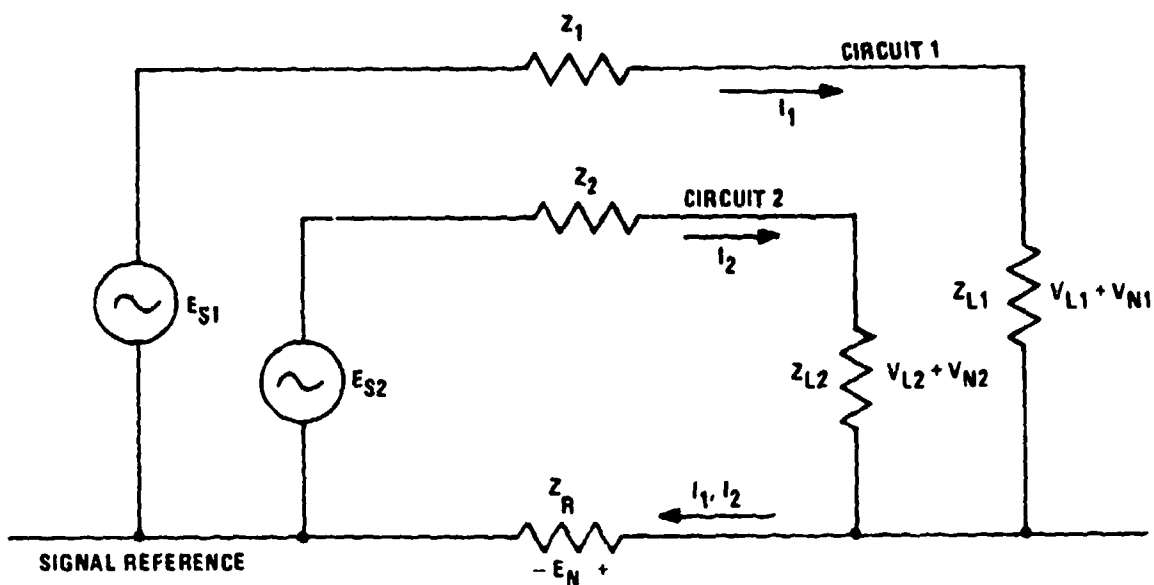


Figure 6-5. Coupling Between Circuits Caused by Common Return Path Impedance

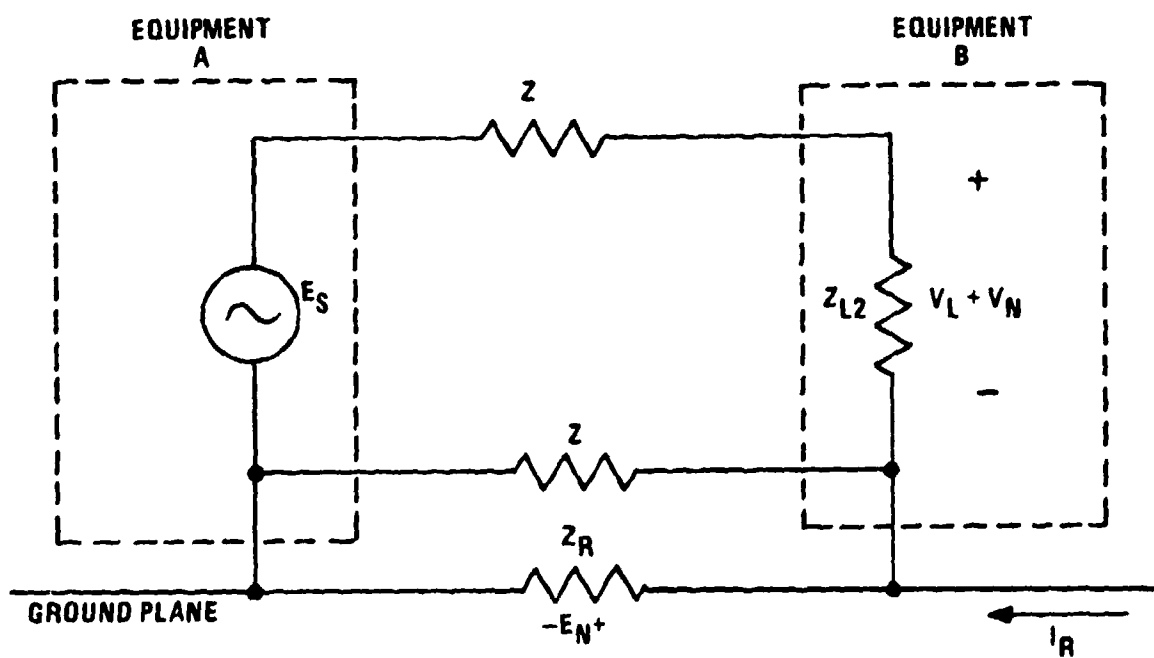


Figure 6-6. Conductive Coupling of Extraneous Noise into Equipment Interconnecting Cables

6.2.2.2 Inductive Coupling. The magnetic field surrounding a long, straight, current-carrying wire as shown in Figure 6-7 is the means for inductive coupling. This field can be determined from Ampere's law (6-1):

$$\oint \underline{H} \cdot d\underline{l} = i(t), \quad (6-1)$$

where  $\underline{H}$  is the magnetic field strength and  $d\underline{l}$  is a small element of length along the path of integration (any closed loop around the current  $i(t)$ ). Choosing a circle of radius  $r$  for the integration path in Equation 6-1 allows one to derive an expression for the magnetic field:

$$H = \frac{i(t)}{2\pi r} \text{ Amperes/meter.} \quad (6-2)$$

Thus the magnetic field strength surrounding a long straight wire carrying a current  $i$  is inversely proportional to the distance  $r$  from the wire, i.e.,  $H$  decreases as the distance from the wire increases.

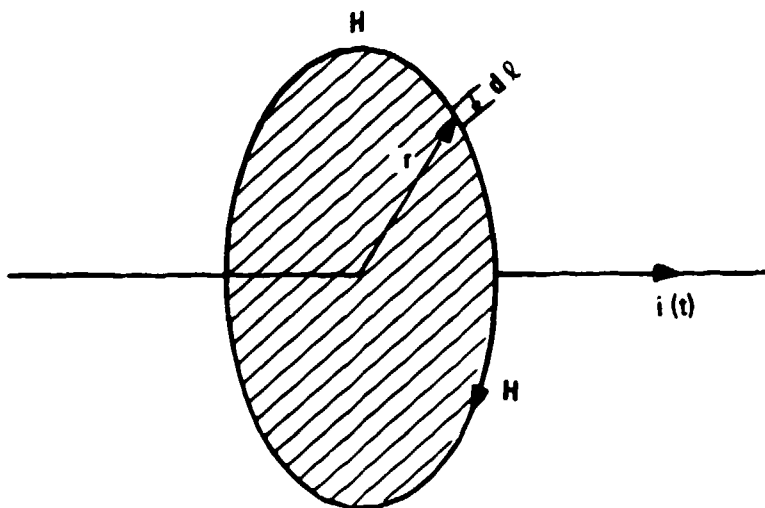


Figure 6-7. Magnetic Field Surrounding a Current-Carrying Conductor

This magnetic field will induce a voltage into a nearby signal circuit loop as illustrated in Figure 6-8. According to Faraday's law (6-1), the induced voltage is

$$v_i(t) = - \frac{d\psi}{dt} \quad \text{volts,} \quad (6-3)$$

where  $v_i(t)$  is the induced voltage and  $\psi$  is the total magnetic flux linking the susceptible circuit loop. This magnetic flux is given by

$$\begin{aligned} \psi &= \int_S \underline{B} \cdot d\underline{s}, \\ &= \mu \int_S \underline{H} \cdot d\underline{s} \quad \text{webers,} \end{aligned} \quad (6-4)$$

where  $\underline{B} = \mu \underline{H}$  is the magnetic flux density,  $\mu$  is the absolute permeability of the medium, and  $d\underline{s}$  is a small element of the loop area. Substituting Equation 6-2 into Equation 6-4 and integrating over the area of the loop in Figure 6-8 gives

$$\psi = \frac{\mu \ell i(t)}{2\pi} \ln \left( \frac{r_2}{r_1} \right) \quad \text{webers,} \quad (6-5)$$

where  $r_1$  and  $r_2$  are the distances from  $i(t)$  to the two sides of the loop which are parallel to  $i(t)$ , and  $\ell$  is the length of each of these sides (in meters). This equation gives the total magnetic flux linking a susceptible

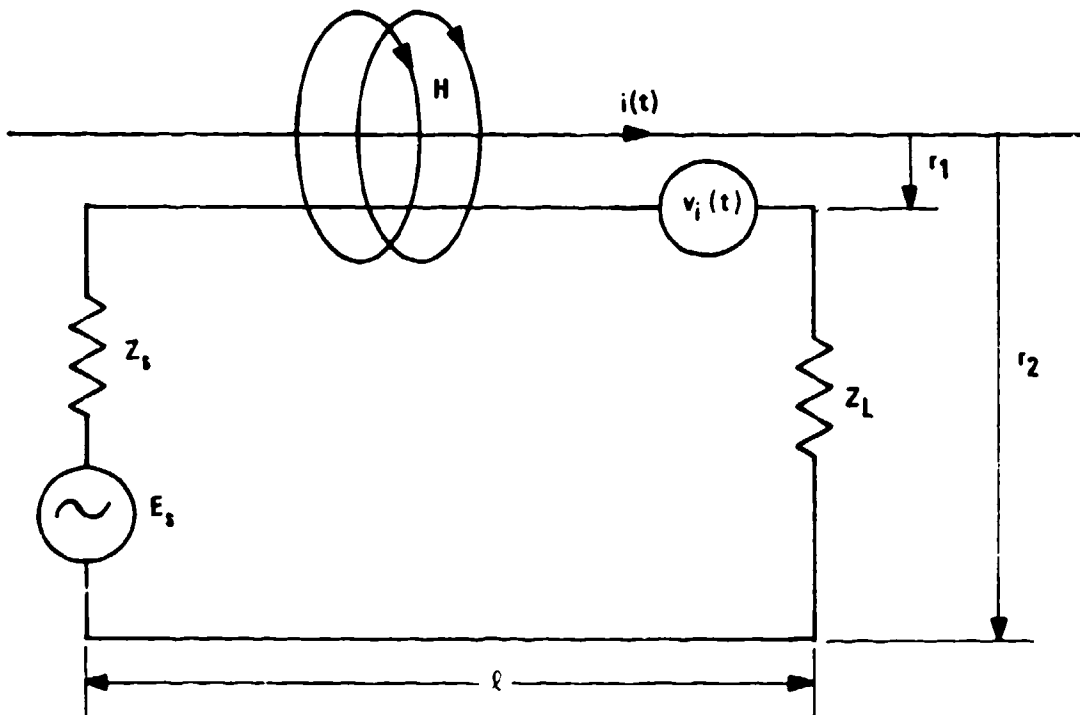


Figure 6-8. Illustration of Inductive Coupling

circuit loop in terms of the current flowing in a nearby conductor parallel to the sides of the loop. Substituting the total magnetic flux from Equation 6-5 into Equation 6-3 gives the voltage  $v_i(t)$ , induced in the circuit loop:

$$v_i(t) = \frac{-\mu \ell}{2\pi} \frac{di}{dt} \ln \left( \frac{r_2}{r_1} \right) \text{ volts.} \quad (6-6)$$

In free space,  $\mu = 4\pi \times 10^{-7}$  henrys/meter, and Equation 6-6 reduces to

$$v_i(t) = -2 \times 10^{-7} \ell \frac{di}{dt} \ln \left( \frac{r_2}{r_1} \right) \text{ volts.} \quad (6-7)$$

If the interference current in the nearby conductor is sinusoidal, i.e.,

$$i(t) = I_m \cos(\omega t + \theta), \quad (6-8)$$

then the maximum value of  $di/dt$  is

$$\left| \frac{di}{dt} \right|_{\max} = \omega I_m = 2\pi f I_m, \quad (6-9)$$

and the maximum value of the induced voltage in Equation 6-6 is

$$\left| v_i \right|_{\max} = \mu \ell f I_m \ln \frac{r_2}{r_1}, \quad (6-10)$$

where  $f$  is the frequency of  $i(t)$ .

In Equations 6-6 and 6-7, the induced voltage  $v_i(t)$  in a circuit loop with sides parallel to a current  $i(t)$  is expressed in terms of the dimensions of the loop, the distance of the loop from the current  $i(t)$  and time rate of change of  $i(t)$ . As can be seen from these equations, the induced voltage in a susceptible circuit loop increases with an increase in the loop area (either an increase in  $\ell$  or  $r_2$  or both), the frequency,  $f$ , or the time rate of change,  $di/dt$ , of the interfering sources, and increases with a decrease in the distance,  $r_1$ , between the interfering source and the loop.

The preceding equations indicate that the induced voltage is independent of the impedance of the susceptible circuit loop; i.e., the amplitude of the induced voltage is the same in a high impedance circuit as it is in a low impedance circuit. The desired signal voltages in low impedance circuits, however, are generally much lower than in high impedance circuits. Therefore, in low impedance circuits the induced voltage can be high relative to the signal voltage and thus more likely to produce significant interference in the circuit load. In high impedance circuits the same induced voltage may be small compared to the circuit signal voltages and thus not create any problems. For these reasons, low impedance circuits are usually more susceptible to inductive coupling than are high impedance circuits.

**6.2.2.3 Capacitive Coupling.** When signal conductors of two circuits are near each other as shown in Figure 6-9, a capacitance,  $C_c$  exists between the conductors. The value of this capacitance is a function of the geometry of the signal lines. For parallel wires,  $C_c$  is given by (6-2).

$$C_c = \frac{1.21}{\log\left(\frac{2r_c}{d}\right)} \times 10^{-11} \text{ farads/meter,} \quad (6-11)$$

where  $r_c$  is the distance between the two lines and  $d$  is the diameter of the wires. In a similar manner, a capacitance exists between each signal line and its return. If the signal line is parallel to its return, these capacitances can be calculated using Equation 6-11 by replacing  $r_c$  with  $r_1$  and with  $r_2$  (see Figure 6-9).

The interference source voltage,  $v_s(t)$  produces a current flow through the mutual capacitance,  $C_c$ , between the two signal conductors and develops an induced voltage,  $v_i(t)$ , in the susceptible circuit. The equivalent circuit for Figure 6-9 is given in Figure 6-10(a) where the parallel combination of  $Z_{S2}$  and  $Z_{L2}$  has been replaced by the equivalent impedance

$$Z_2 = \frac{Z_{S2}Z_{L2}}{Z_{S2} + Z_{L2}}, \quad (6-12)$$

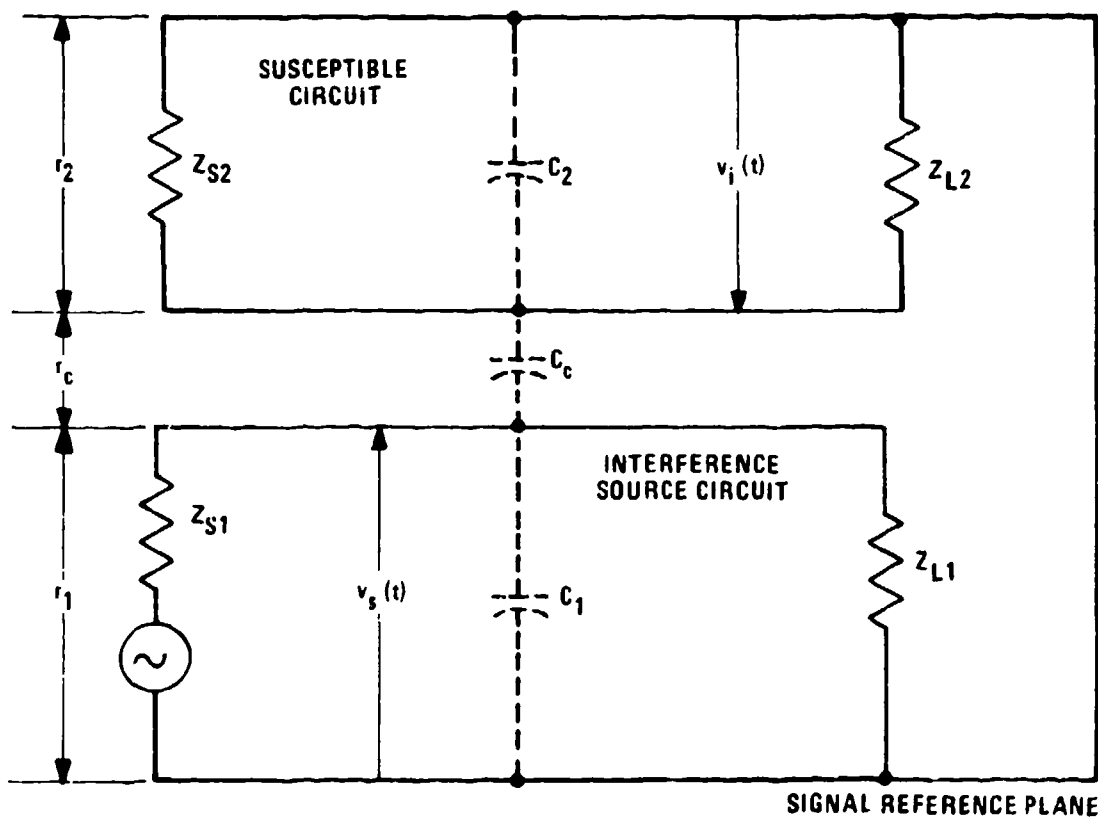


Figure 6-9. Illustration of Capacitive Coupling

and time-varying voltages have been replaced by their ac steady state phasors. The induced voltage (ac steady state assumed) in the susceptible circuit is

$$V_i = \frac{Z_{C2}Z_2/(Z_{C2} + Z_2)}{Z_{CC} + Z_{C2}Z_2/(Z_{C2} + Z_2)} V_s, \quad (6-13)$$

where

$$Z_{C2} = \frac{1}{j\omega C_2} \quad (6-14)$$

and

$$Z_{CC} = \frac{1}{j\omega C_c} \quad (6-15)$$

Substitution of Equations 6-14 and 6-15 into Equation 6-13 yields

$$V_i = \frac{j\omega Z_2 C_c}{1 + j\omega Z_2 (C_2 + C_c)} V_s, \quad (6-16)$$

where  $\omega = 2\pi f$ . If  $j\omega Z_2 C_c \ll 1$ , which is generally true at lower frequencies, the equivalent circuit of Figure 6-10(b) is applicable and

$$V_i = j2\pi f Z_2 C_c V_s. \quad (6-17)$$

At higher frequencies, the equivalent circuit of Figure 6-10(c) is applicable and

$$V_i = \frac{C_c}{C_2 + C_c} V_s. \quad (6-18)$$

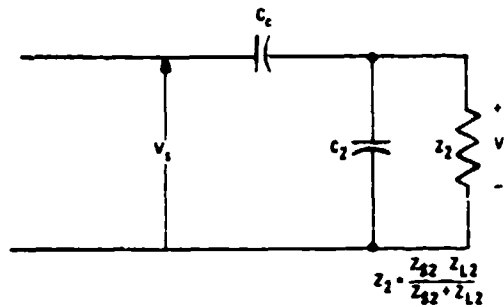
These equations illustrate the induced voltage,  $V_i$ , which is capacitively coupled into a susceptible signal circuit from a nearby signal conductor, is dependent on the amplitude and frequency of the interference source voltage,  $V_s$ , the values of the coupling capacitance,  $C_c$ , the stray capacitance in the susceptible circuit,  $C_2$ , and on the magnitude of the impedance of the susceptible circuit. At low frequencies, Equation 6-17 indicates that the induced voltage increases with either an increase in the coupling capacitance or an increase in the impedance of the susceptible loop. Similarly, at high frequencies the induced voltage as given in Equation 6-18 increases with either an increase in the coupling capacitance or a decrease in the stray capacitance of the susceptible circuit. It should also be noted that the value of the interference source voltage,  $V_s$ , depends upon the stray capacitance in the interference source circuit,  $C_1$  in Figure 6-9.

If the impedance of the susceptible circuit is totally resistive, i.e., if  $Z_2 = R_2$ , the induced voltage is given by Equation 6-16 as

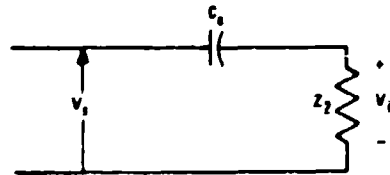
$$V_I = \frac{j2\pi f R_2 C_c}{1 + j2\pi f R_2 (C_2 + C_c)} V_s, \quad (6-19)$$

and the magnitude of the ratio of the induced voltage to the interference (sinusoidal) voltage is

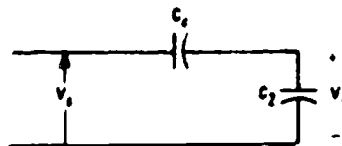
$$\left| \frac{V_I}{V_s} \right| = \sqrt{\frac{(2\pi f R_2 C_c)^2}{1 + (2\pi f R_2 (C_2 + C_c))^2}},$$



(a) TRUE EQUIVALENT CIRCUIT



(b) LOW FREQUENCY APPROXIMATION



(c) HIGH FREQUENCY APPROXIMATION

Figure 6-10. Equivalent Circuit of Network in Figure 6-9



$$= \frac{C_c}{C_2 + C_c} \sqrt{\frac{R_2^2}{R_2^2 + \left[ \frac{1}{2\pi f(C_2 + C_c)} \right]^2}} \quad (6-20)$$

This ratio increases almost linearly with  $R_2$  until  $R_2$  approaches the value  $\frac{1}{2\pi f(C_2 + C_c)}$ ; i.e., the reactance of  $C_2$  and  $C_c$  in parallel. For larger values of  $R_2$ , the ratio asymptotically approaches  $C_c/(C_2 + C_c)$ . The behavior of this voltage ratio with frequency is illustrated in Figure 6-11. The ratio is zero at dc and asymptotically approaches  $C_c/(C_2 + C_c)$  as the frequency is increased. Equation 6-20 and Figure 6-11 illustrate again that the voltage capacitively coupled into the susceptible circuit increases with an increase in the total resistance of the circuit and with an increase in frequency. Resonances can occur and change the amount of capacitive coupling if the impedance of the susceptible circuit contains inductive reactance, but such resonances usually only produce noticeable effects at higher frequencies.

#### 6.2.2.4 Far-Field Coupling.

Radiation is the means by which energy escapes from a conductor and propagates into space. The conductor does not have to be specifically designed to radiate energy; it may be any current carrying conductor; e.g., a signal line, a power line, or even a ground lead.

Algebraic expressions for the electromagnetic fields surrounding a current carrying conductor are usually expressed as the sum of three terms. Each term is inversely proportional to a power of the distance,  $r$ , from the conductor, i.e., each term is proportional to either  $1/r$ ,  $1/r^2$ , or  $1/r^3$ . Close to the conductor (near field), the  $1/r^2$  and  $1/r^3$  components dominate and the electromagnetic energy oscillates between the space surrounding the conductor and the conductor itself; zero average energy is propagated by the near field terms.

Outside the near field region, the  $1/r$  term predominates. In this far field region, radiated energy that has escaped is propagating away from the "antenna" through space. The mechanism of energy radiation can be visualized (6-3) by considering the finite time required for the electromagnetic fields to propagate between two points in space. Current flows through an antenna at the frequency of the applied signal, and the polarity of the field produced by this current is reversed at this same frequency. When a positive charge is present at one end of the antenna, an equal but, negative charge is present at the other end and an electric field in the vicinity of the antenna will be established between the charges. As the current changes direction, the charges will reverse positions; the electromagnetic field will collapse and be re-established in the opposite direction. If the frequency of the applied signal is low, sufficient time will exist between reversals for practically all the energy stored in the field to be returned to the circuit and very little radiation will occur. If, however, the frequency is high and the charges reverse quickly, a field in the opposite direction is formed near the wire before a substantial amount of the field energy can return to the circuit. This part of the field is thus separated from the antenna and propagates outward through space as an electromagnetic wave.

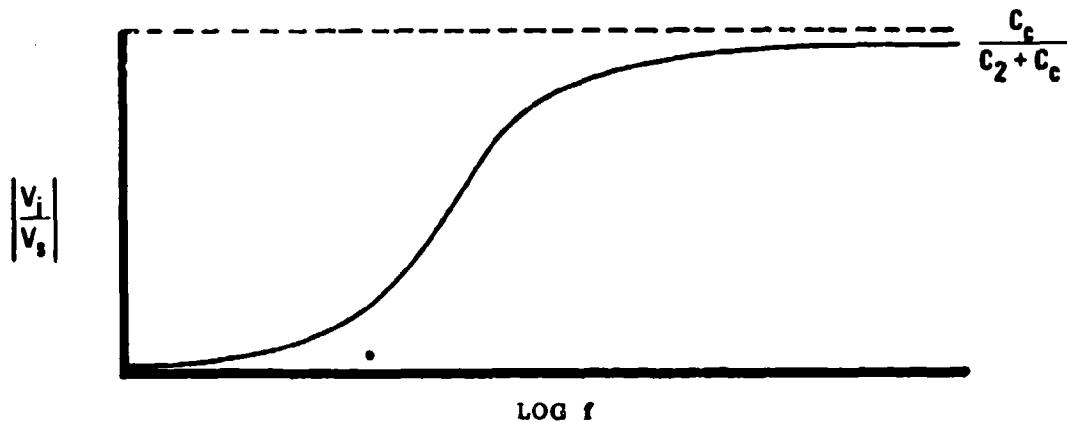


Figure 6-11. Characteristic Voltage Transfer Curve for Capacitive Coupling

This method of visualizing radiation from a wire or antenna is illustrated for a dipole antenna in Figure 6-12. Figure 6-12(a) shows a dipole when the charges are maximum at the ends of the antenna. As the current flow reverses directions and the charges move back toward the center, the electric field lines collapse as in (b). Since the field moves with a finite velocity, there is not enough time for all the field lines to return to the antenna. When the ends of these lines meet at the center of the antenna and the charge on the antenna is zero, the field lines that have not collapsed will close on themselves and continue to exist as closed loops as illustrated in (c). The antenna charges move in the opposite direction as shown in (d), and the oppositely directed electric field pushes away the previously detached loop as shown in (e). This procedure continues with the fields in the opposite direction, and a cycle is completed when the fields near the antenna return to their original state. These cycles repeat at the frequency of the applied signal, and an electromagnetic field propagates outward from the antenna at the speed of light. Although only the electric field is illustrated, there is an associated magnetic field in accordance with Maxwell's equations (6-1). The magnetic field consists of concentric circles surrounding the antenna and expanding radially as the electric field propagates outward. These outward propagating electric and magnetic fields represent energy flowing away from the antenna. Therefore, the antenna radiates energy into the surrounding space.

In a reciprocal manner, wires and conductors located in a radiated field have currents induced in them and act as receiving antennas for incident electromagnetic energy. These induced currents in the wires can then be conducted into associated signal circuitry as interference (see Section 6.2.1). The amplitude of the resulting interference depends on the strength of the electromagnetic field in the vicinity of the wire and on the efficiency of the wire as an antenna. The strength of this field is a function of the distance from the radiating wire, the efficiency of the radiating wire as an antenna, and the amplitude and frequency of the signal on the radiating wire. The efficiency of a wire or other conductor as either a receiving or a radiating antenna is a function of the length of the wire relative to the wavelength of the signal.

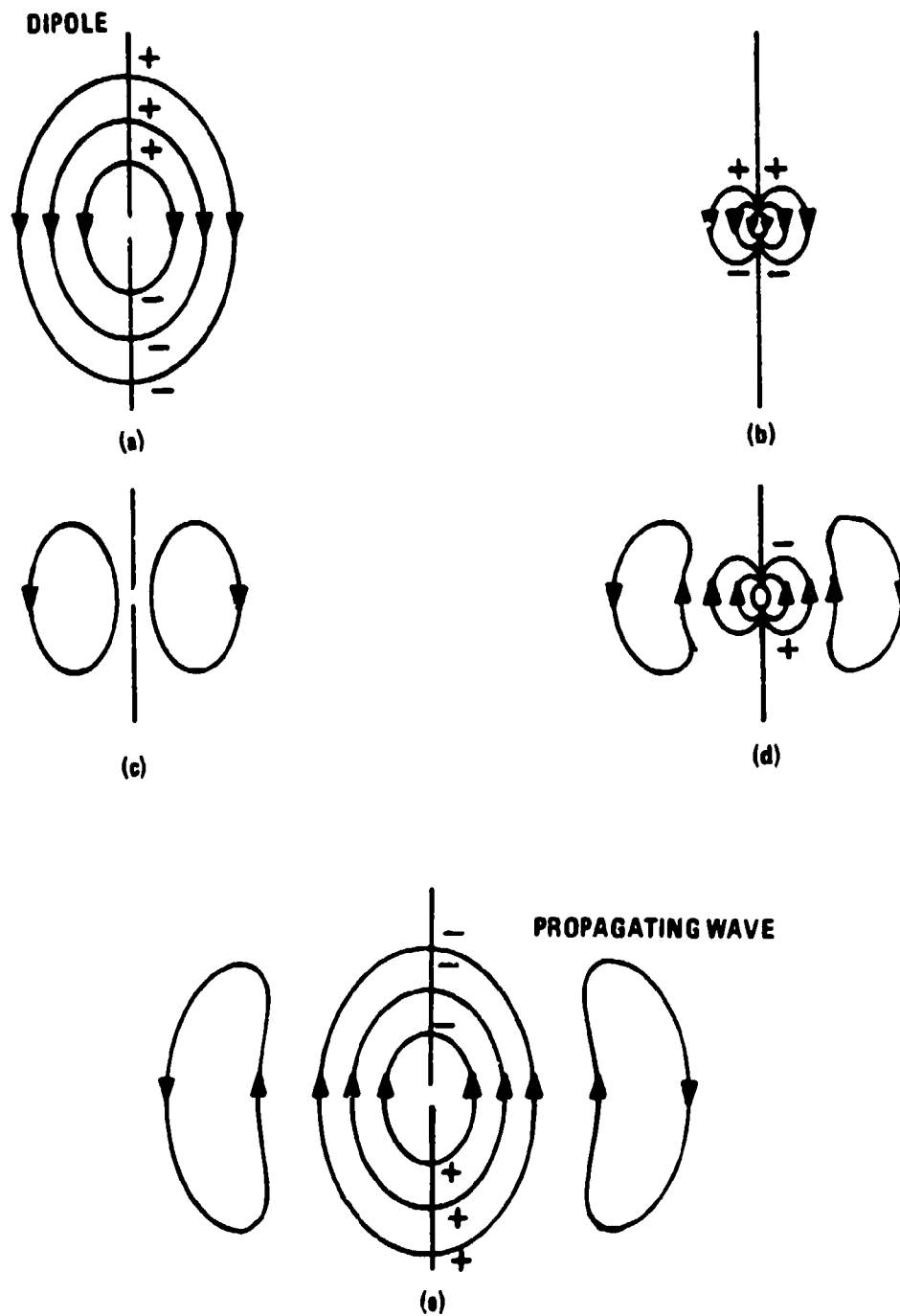


Figure 6-12. Electric Field Patterns in the Vicinity of a Radiating Dipole

One way of evaluating the efficiency of a wire as an antenna is to compare its radiation resistance with the radiation resistance of a quarter-wave ( $\lambda/4$ ) antenna. The radiation resistance of an antenna is the resistance which would consume the same amount of power as is radiated by the antenna. Thus the radiation resistance is a direct measure of the energy radiated from the antenna. A monopole antenna one-quarter of a wavelength long has a radiation resistance of 36.5 ohms (6-4). An antenna which transmits or receives ten percent or less of the energy that would be transmitted or received by a  $\lambda/4$  monopole can be considered relatively inefficient. Thus an inefficient antenna would exhibit a radiation resistance of 3.65 ohms or less. Monopoles of length  $\lambda/11$  meet this criterion (6-4). Greater convenience in calculations results if  $\lambda/10$  is chosen instead of  $\lambda/11$ . Thus  $\lambda/10$  is chosen to represent the length below which a conductor does not perform effectively as an antenna.

### 6.3 COMMON-MODE NOISE.

Common-mode noise is an unwanted noise voltage which appears identically on both sides of a signal line when measured from the system ground or common point. It, like normal-mode noise, can be caused by resistive coupling, capacitive coupling, or magnetic coupling from the unwanted source. In addition, many measuring transducers intentionally have a dc or ac common-mode voltage present on both output lines, the presence of which is necessary for proper operation of the transducer. Although not a noise source, these common-mode voltages require careful design and use of data and instrumentation amplifiers to prevent interference with the desired signal components.

The source of most common-mode noise is resistive coupling between separate ground points in a circuit or system. A simple example of this is illustrated in Figure 6-13. An oscilloscope probe is used to couple a signal from some point in a circuit to the oscilloscope terminals. The probe ground is connected to circuit ground which is in turn referenced through the facility ground system. Since there are generally currents flowing in the facility ground system (these are primarily at the 60 Hz power line frequency), it follows that the ground reference potential for the circuit is different from that for the oscilloscope. This difference in potential is produced by the flow of the stray ground currents through the impedance of the facility ground system. Thus, both the ground reference for the circuit and the signal point in the circuit have identical noise voltages impressed on them with respect to the ground reference for the oscilloscope. This noise is called common-mode noise by virtue of the fact that is common to all points in the circuit, including the circuit ground. Not only do these noise sources introduce measurement errors but they also produce interference between interconnected equipments.

Resistively coupled common-mode noise can also occur in a single equipment rather than between equipments. The coupling arises from multiple signal currents and power frequency currents flowing in a common ground lead, chassis, or ground plane.

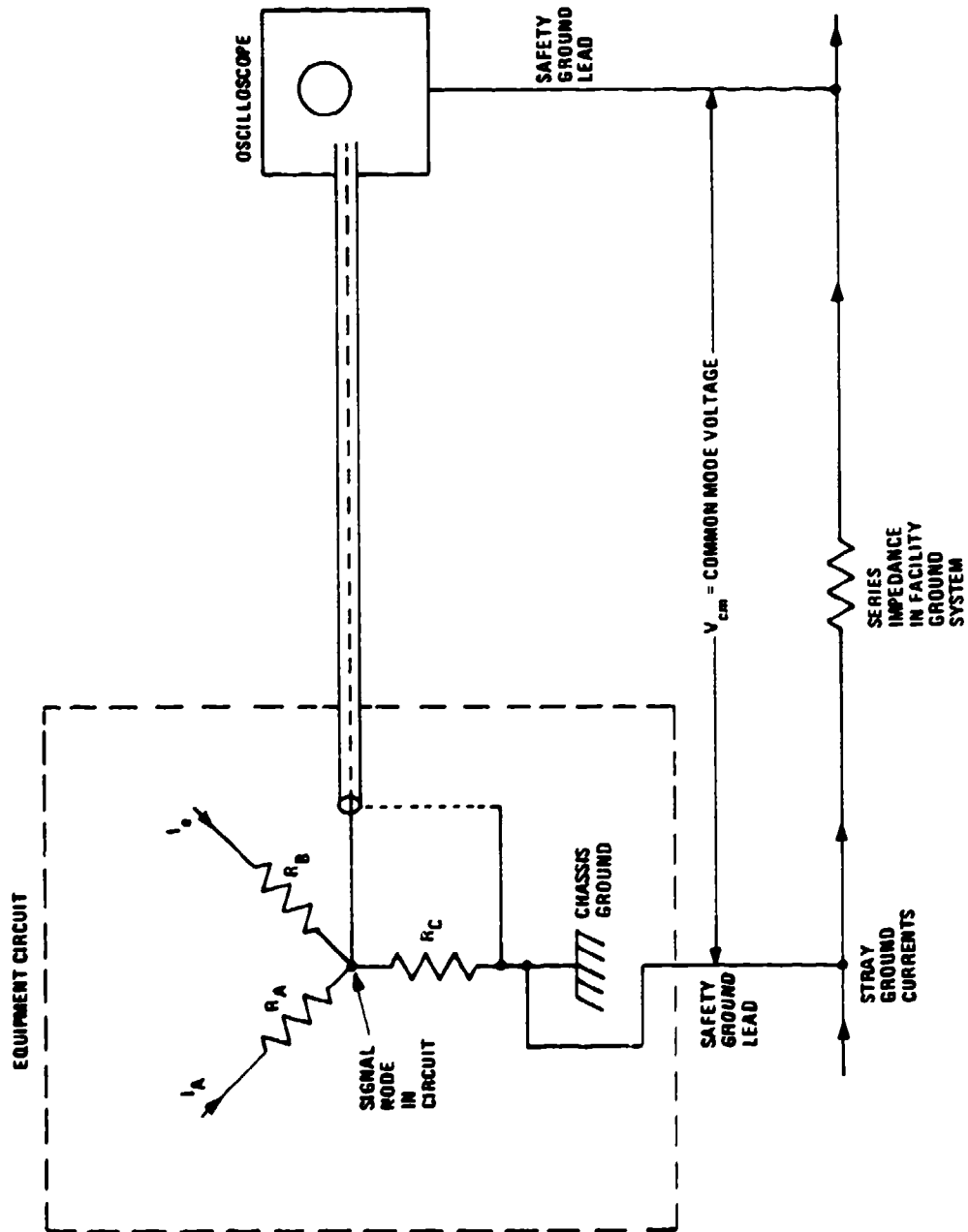


Figure 6-13. Illustration of Conductively-Coupled Common-Mode Noise

### 6.3.1 Basic Theory of Common-Mode Coupling.

The mechanism of common-mode coupling can be explained with reference to Figure 6-14. In this figure,  $V_s$  represents some signal voltage from an unbalanced source, i.e., the output signal of some transducer or measuring amplifier, and  $R_s$  is the output impedance of this source. The source is connected to the input terminals of some electronic device which is modeled as a two-terminal pair amplifier in the figure.  $R_1$  and  $R_2$  are the series resistances in the interconnecting cables between the source and amplifiers. The voltage source  $V_{cm}$  with output resistance  $R_{cm}$  represents a common-mode noise voltage source which causes the signal source to be at some voltage when measured with respect to the ground reference of the amplifier output. In Figure 6-14, the impedances  $Z_1$  and  $Z_2$  represent the input impedances of the two amplifier terminals. In a differential amplifier, these impedances are normally very high, however, in a single ended amplifier, one is high and the other is very low since it is tied directly to the ground reference terminal.

The analysis of the circuit in Figure 6-14 is complicated enough to make it difficult to reach conclusions without excessive algebra. Normally,  $R_{cm}$  is small and can be neglected. With this approximation, it can be shown that the output voltage of the amplifier is given by

$$\begin{aligned}
 V_o &= KV_x \\
 &= K \left( \frac{Z_1}{R_s + R_1 + Z_1} - \frac{Z_2}{R_2 + Z_2} \right) V_{cm} \\
 &\quad + \left( \frac{KZ_1}{R_s + R_1 + Z_1} \right) V_s,
 \end{aligned} \tag{6-21}$$

where  $K$  is the voltage gain of the amplifier.

There are two signal contributions to the output signal  $V_o$  in Equation 6-21: the desired signal and the undesired common-mode noise. There are three ways in which the common-mode noise term can be reduced. These are as follows:

- a. Decrease  $V_{cm}$  - By decreasing  $V_{cm}$ , the common-mode noise voltage at the output terminals decreases proportionally.

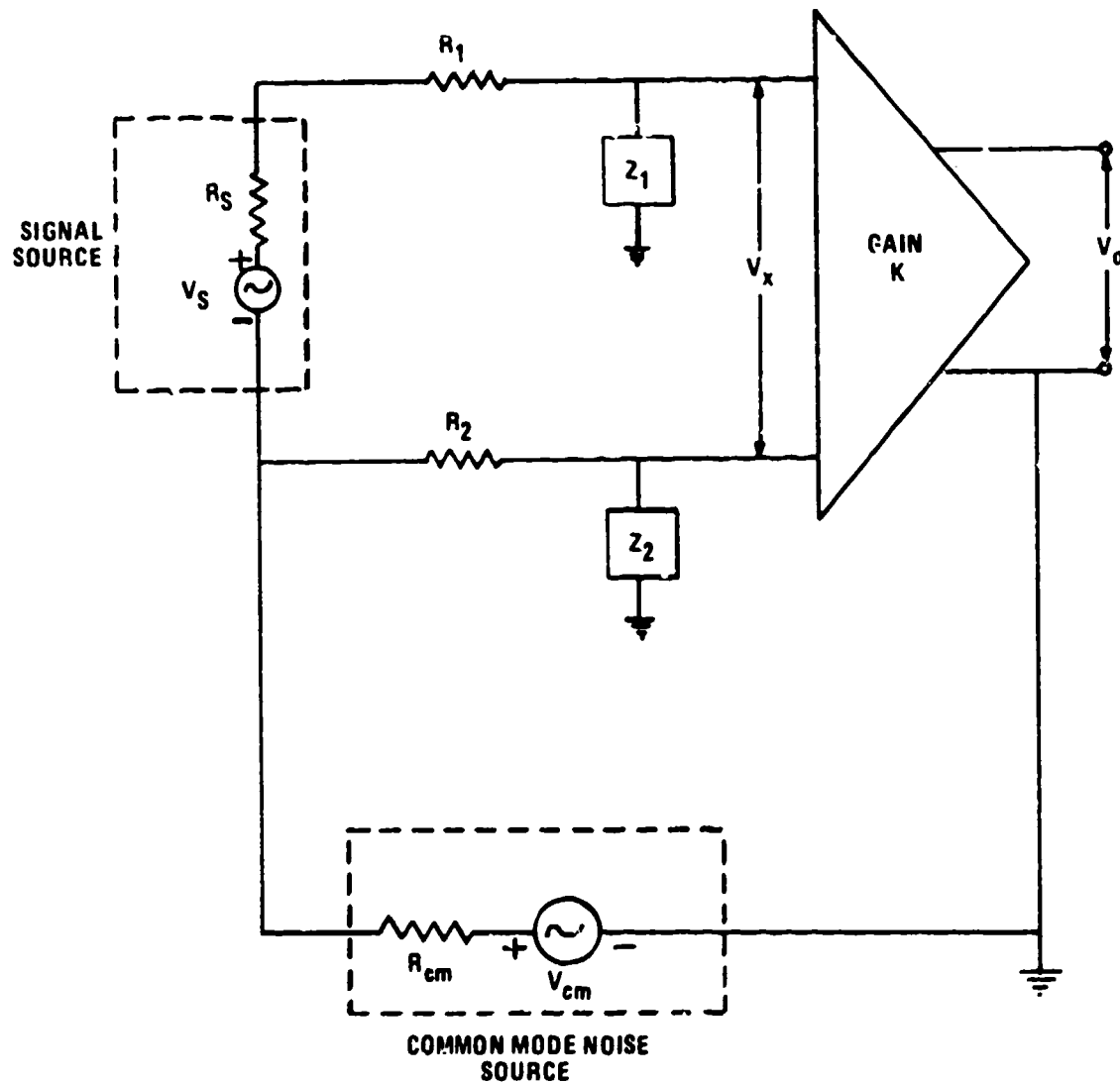


Figure 6-14. Common-Mode Noise in Unbalanced Systems

- b. Balance the Two Amplifier Inputs - If  $R_1$  and  $R_2$  are manipulated such that

$$\frac{Z_1}{R_s + R_1 + Z_1} = \frac{Z_2}{R_2 + Z_2}, \quad (6-22)$$

the common-mode noise voltage at the amplifier output terminals can be made to vanish.

- c. Increase  $Z_1$  and  $Z_2$  - If  $Z_1$  is sufficiently large compared to  $R_s + R_1$ , and  $Z_2$  is sufficiently large compared to  $R_2$ , then the common-mode noise voltage at the amplifier output terminals will be diminished. This approach normally requires a differential amplifier with carefully shielded input signal lines.

In the case of balanced signal sources or transducers, the basic circuit and equations differ from those given in Figure 6-14 and by Equation 6-21. Figure 6-15 shows a balanced source with an output voltage  $V_s$  and output resistance  $R_s$  connected to the two inputs of an amplifier. In this case, the center tap of the source is connected to the ground reference terminal. As before, if it is assumed that  $R_{cm}$  is small, it can be shown that  $V_o$  is given by

$$\begin{aligned} V_o &= KV_x, \\ &= \left( \frac{Z_1}{\frac{R_s}{2} + R_1 + Z_1} - \frac{Z_2}{\frac{R_s}{2} + R_2 + Z_2} \right) V_{cm} \\ &\quad + K \left( \frac{Z_1}{\frac{R_s}{2} + R_1 + Z_1} + \frac{Z_2}{\frac{R_s}{2} + R_2 + Z_2} \right) \frac{V_s}{2} \end{aligned} \quad (6-23)$$

The same conclusions regarding the minimization of the common-mode noise component at the amplifier output apply in this circuit as for the unbalanced source. However, in this case the amplifier must have a differential input stage. Otherwise, one-half of the source would be shorted out. In Figure 6-14, the amplifier can have single-ended or differential inputs.

The common-mode rejection (CMR) ratio of an amplifier is the gain of the amplifier ( $K$ ) multiplied by the common-mode noise voltage ( $V_{cm}$ ) and divided by the amplifier output due to  $V_{cm}$ . The CMR ratio describes a circuit's ability to avoid converting common-mode noise to normal-mode noise. Expressed as a positive quantity, the CMR ratio is given by

$$CMR = \left| \frac{KV_{cm}}{V_o} \right| \quad V_s = 0 \quad (6-24)$$



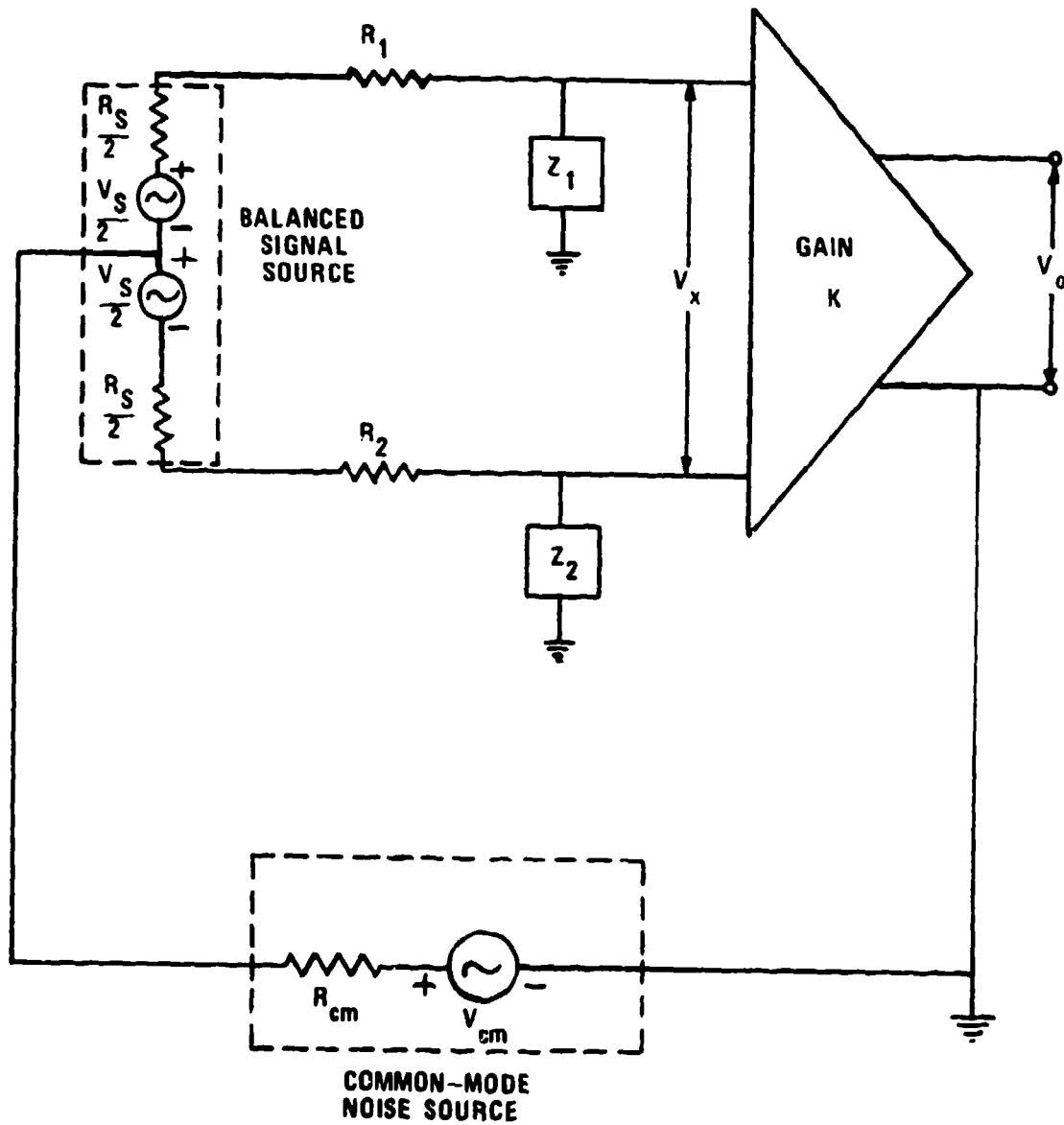


Figure 6-15. Common-Mode Noise in Balanced Systems

Ideally, the CMR of an amplifier should be infinite, or as large as possible. Under the worse case conditions,  $\text{CMR} = 1$ . As it is defined, the CMR conveys a measure of how well the amplifier can reject a common-mode noise signal at its input. Typical values for a good differential amplifier with balanced input impedances are in the vicinity of  $\text{CMR} = 1000$ . Often this is expressed in decibels which, in this case, would be  $\text{CMR} = 60 \text{ dB}$ .

The CMR for the amplifier in Figure 6-14 is easily derived from Equation 6-21 to be

$$\text{CMR} = \left| \frac{1}{\frac{Z_1}{R_s + R_1 + Z_1} - \frac{Z_2}{R_2 + Z_2}} \right| \quad (6-25)$$

**6.3.2 Differential Amplifier.** A differential amplifier is designed to make  $Z_2$  large compared to  $R_2$  and  $Z_1$  large compared to  $R_1 + R_s$ . Since  $Z_1$  and  $Z_2$  are normally functions of frequency, it can be seen that the CMR will also be a function of frequency. Typically  $Z_1$  and  $Z_2$  are resistors shunted by capacitors. Thus, it can be seen that the CMR will inevitably decrease with increasing frequency when the capacitive reactance becomes smaller than the resistor. Consequently, a high CMR is difficult to achieve at high frequencies.

**6.4 MINIMIZATION TECHNIQUES.** Signal interaction, i.e., interference, can be minimized by reducing the coupling between the signal systems by modifying the signal systems in such a manner that interaction between the systems does not produce interference in either one, by eliminating the source of the interference, and by filtering the interference out of the susceptible signal system.

**6.4.1 Reduction of Coupling.** The techniques for reducing coupling include minimizing the impedance of the reference plane, increasing the spatial separation between the signal systems, shielding the systems from each other, reducing the loop area of each signal system, and balancing the signal lines in each system.

**6.4.1.1 Reference Plane Impedance Minimization.** Minimizing the impedance of the signal reference plane lowers the potential difference between any two points in the reference plane, thereby reducing the conductive coupling of interference in susceptible circuits referenced to these points. The impedance of the signal reference plane is reduced by minimizing both the resistance ( $R$ ) and the series reactance ( $X$ ) of the conductors forming the reference plane. The resistance decreases with a decrease in either the length of the conductors or the signal frequency (because of skin effect - see Section 5.2.2.1) and with an increase in conductor cross-sectional area. The reactance also decreases with a decrease in the signal frequency and with a decrease in the inductance of the conductors; the inductance is a function of both the conductor length and cross-sectional area. The impedance of the signal reference plane can be reduced by making the reference plane conductors as short as possible and by using conductors with cross-sectional areas as large as practical. The overall impedance of the signal reference plane also depends upon the establishment of low impedance bonds between ground conductors. (The various aspects of bonding and bond resistance are discussed in Chapter 7.)

**6.4.1.2 Spatial Separation.** Inductive or capacitive coupling can be reduced by increasing the physical distance between signal circuits. As can be seen from Equation 6-6 and Equations 6-11 and 6-16, increasing the separation between the interfering circuit and the susceptible circuit exponentially decreases the voltage coupled into the susceptible circuit.

**6.4.1.3 Reduction of Circuit Loop Area.** Reducing the loop area of either the interference source circuit or the susceptible circuit will decrease the inductive coupling between the circuits. Equation 6-6 shows that the inductively coupled voltage can be minimized by reducing the length ( $l$ ) or the width ( $r_2 - r_1$ ) of the susceptible circuit. This width can be minimized by running the signal return adjacent to the signal conductor and, hence, reducing the loop area of the susceptible circuit. A preferable approach is to twist the signal conductor with its return. The use of twisted wires reduces the inductively coupled voltages since the voltage induced in each small twist area is approximately equal and opposite to the voltage induced in the adjacent twist area.

**6.4.1.4 Shielding.** Another effective means for the reduction of coupling is the use of shields around the circuits and around interconnecting lines. Principles of shielding are presented in Chapter 8.

**6.4.1.5 Balanced Lines.**

In situations where signal circuits must be grounded at both the source and the load, and hence, establish conductive coupling paths, the use of balanced signal lines and circuits is an effective means of minimizing the conductively coupled interference. In a balanced circuit, the two signal conductors are symmetrical with respect to ground. At equivalent points on the two conductors the desired signal is opposite in polarity and equal in amplitude relative to ground. A common-mode voltage will be in phase and will exhibit equal amplitudes on each conductor and will tend to cancel at the load. The amount of cancellation depends upon the degree to which the two signal lines are balanced relative to ground.

If the signal lines are perfectly balanced, the cancellation would be complete and the coupled interference voltage at the load will be zero. If the source and load are not normally or cannot be operated in a balanced mode, balanced-to-unbalanced transformers or other coupling devices should be used at both the source and load ends of the signal line.

**6.4.2 Alternate Methods.**

Several alternate methods exist for minimizing interference besides the reduction of coupling. The first technique consists of actual circuit modification. For example, the signal frequency of either the interfering source or the susceptible circuit can be changed such that the signals do not interfere with one another. Similarly, the desired signal can be transposed to another frequency range or to a type of signal not affected by the noise. An example of the former is the conversion of the desired signal to VHF/UHF or microwave while an example of the latter is the use of acoustic coupling and electro-optical transmission. Through the use of one of these techniques, the frequency of transmission over that portion of the path susceptible to pickup is such

that the receiving and detection devices do not respond to the noise signal. As another option, the amplitude of the interference source or the sensitivity of the susceptible circuit can be decreased to reduce the interaction between the two circuits. Further, the type of modulation used in one or both circuits can be changed to minimize the interference.

Another technique is the elimination of the interference source. Although this may seem like a trivial solution, it is a valid alternative in many situations. For example, the source of interference may be a rusty joint which can be eliminated by proper bonding.

A third alternative is the use of filters. The majority of interfering signals, even if they are free-space coupled to the signal and power lines, are conductively coupled into the susceptible circuit. The proper application of filters to both the signal and power lines can reduce this coupling.

**6.5 FACILITY AND EQUIPMENT REQUIREMENTS.** The interference rejection principles identified in this chapter are responsible, in part, for many of the recommendations contained elsewhere in this volume and in Volume II. For example, intersite or interbuilding common-mode noise voltages in the earth contribute to the need for a low resistance of 10 ohms to earth at each facility. Even a resistance to earth of as low as 10 ohms may not, however, alleviate all common mode noise on a data cable connecting two separate locations or buildings. While a low resistance may help, there will always be potential differences between any two rods in the ground. The use of shielded, balanced twisted pair for all lower frequency equipment interfaces recommended in Volume II, is intended to provide additional common-mode rejection to those unavoidable noise voltages which exist in any facility. This is not to say that the sources of noise in a facility cannot be controlled. In fact, much can be done by equalizing the load between the phases of the ac distribution system; by insuring that the neutral is grounded only at the service disconnecting means as recommended in Volume II; and by limiting the quantity of leakage current from power line filter capacitors by using the smallest acceptable value of capacitance or by sharing common filtered lines with several pieces of equipment.

#### **6.6 REFERENCES.**

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- 6-3. L.V. Blake, Antennas, John Wiley and Sons, Inc, New York NY, (1966).
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## CHAPTER 7

## BONDING

7.1 DEFINITION OF BONDING. As used in these Volumes, bonding refers to the process by which a low impedance path for the flow of an electric current is established between two metallic objects. Other types of bonding which involve simply the physical attachment of one substance or object to another through various mechanical or chemical means are not treated.

7.2 PURPOSES OF BONDING.

In any realistic electronic system, whether it be only one piece of equipment or an entire facility, numerous interconnections between metallic objects must be made in order to provide electric power, minimize electric shock hazards, provide lightning protection, establish references for electronic signals, etc. Ideally, each of these interconnections should be made so that the mechanical and electrical properties of the path are determined by the connected members and not by the interconnection junction. Further, the joint must maintain its properties over an extended period of time in order to prevent progressive degradation of the degree of performance initially established by the interconnection. Bonding is concerned with those techniques and procedures necessary to achieve a mechanically strong, low impedance interconnection between metal objects and to prevent the path thus established from subsequent deterioration through corrosion or mechanical looseness.

In terms of the results to be achieved, bonding is necessary for the:

- a. protection of equipment and personnel from the hazards of lightning discharges,
- b. establishment of fault current return paths,
- c. establishment of homogeneous and stable paths for signal currents,
- d. minimization of rf potentials on enclosures and housings,
- e. protection of personnel from shock hazards arising from accidental power grounds, and
- f. prevention of static charge accumulation.

With proper design and implementation, bonds minimize differences in potential between points within the fault protection, signal reference, shielding, and lightning protection networks of an electronic system. Poor bonds, however, lead to a variety of hazardous and interference-producing situations. For example, loose connections in ac power lines can produce unacceptable voltage drops at the load, and the heat generated by the load current through the increased resistance of the poor joint can be sufficient to damage the insulation of the wires which may produce a power line fault or develop a fire hazard or both. Loose or high impedance joints in

signal lines are particularly annoying because of intermittent signal behavior such as decreases in signal amplitude, increases in noise level, or both. Poor joints in lightning protection networks can be particularly dangerous. The high current of a lightning discharge may generate several thousand volts across a poor joint. Arcs produced thereby present both a fire and explosion hazard and may possibly be a source of interference to equipments. The additional voltage developed across the joint also increases the likelihood of flashover occurring to objects in the vicinity of the discharge path.

A degradation in system performance from high noise levels is frequently traceable to poorly bonded joints in circuit returns and signal referencing networks. As noted previously, the reference network provides low impedance paths for potentially incompatible signals. Poor connections between elements of the reference network increase the resistance of the current paths. The voltages developed by the currents flowing through these resistances prevent circuit and equipment signal references from being at the same reference potential. When such circuits and equipments are interconnected, the voltage differential represents an unwanted signal within the system.

Bonding is also important to the performance of other interference control measures. For example, adequate bonding of connector shells to equipment enclosures is essential to the maintenance of the integrity of cable shields and to the retention of the low loss transmission properties of the cables. The careful bonding of seams and joints in electromagnetic shields is essential to the achievement of a high degree of shielding effectiveness. Interference reduction components and devices also must be well bonded for optimum performance. Consider a typical power line filter like that shown in Figure 7-1. If the return side of the filter (usually the housing) is inadequately bonded to the ground reference plane (typically the equipment case or rack), the bond impedance  $Z_B$  may be high enough to impair the filter's performance. The filter as shown is a low pass filter intended to remove high frequency interference components from the power lines of equipment. The filter achieves its goal in part by the fact that the reactance,  $X_C$ , of the shunt capacitors is low at the frequency of the interference. Interfering signals present on the ac line are shunted to ground along Path 1 and thus do not reach the load. If  $Z_B$  is high relative to  $X_C$ , however, interference currents will follow Path 2 to the load and the effectiveness of the filter is compromised.

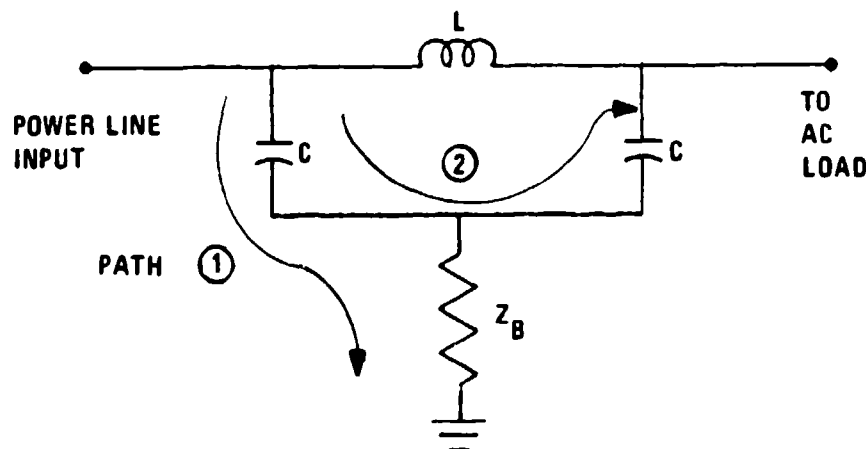


Figure 7-1. Effects of Poor Bonding on the Performance of a Power Line Filter

If a joint in a current path is not securely made or works loose through vibration, it can behave like a set of intermittent contacts. Even if the current through the joint is at dc or at the ac power frequency, the sparking which occurs may generate interference signals with frequency components up to several hundred megahertz.

Poor bonds in the presence of high level rf fields, such as those in the immediate vicinity of high powered transmitters, can produce a particularly troublesome type of interference. Poorly bonded joints have been shown to generate cross modulation and other mix products when irradiated by two or more high level signals (7-1). Some metal oxides are semiconductors and behave as nonlinear devices to provide the mixing action between the incident signals. Interference thus generated can couple into nearby susceptible equipments.

### 7.3 RESISTANCE CRITERIA.

A primary requirement for effective bonding is that a low resistance path be established between the two joined objects. The resistance of this path must remain low with use and with time. The limiting value of resistance at a particular junction is a function of the current (actual or anticipated) through the path. For example, where the bond serves only to prevent static charge buildup, a very high resistance, i.e., 50 kilohms or higher, is acceptable. Where lightning discharge or heavy fault currents are involved, the path resistance must be very low to minimize heating effects.

Noise minimization requires that path resistances of less than 50 milliohms be achieved. However, noise control rarely ever requires resistances as low as those necessary for fault and lightning currents. Bond resistance based strictly on noise minimization requires information on what magnitude of voltage constitutes an interference threat and the magnitude of the current through the junction. These two factors will be different for every situation.

A bonding resistance of 1 milliohm is considered to indicate that a high quality junction has been achieved. Experience shows that 1 milliohm can be reasonably achieved if surfaces are properly cleaned and adequate pressure is maintained between the mating surfaces. A much lower resistance could provide greater protection against very high currents but could be more difficult to achieve at many common types of bonds such as at connector shells, between pipe sections, etc. However, there is little need to strive for a junction resistance that is appreciably less than the intrinsic resistance of the conductors being joined.

Higher values of resistance tend to relax the bond preparation and assembly requirements. These requirements should be adhered to in the interest of long term reliability. Thus, the imposition of an achievable, yet low, value of 1 milliohm bond resistance ensures that impurities are removed and that sufficient surface contact area is provided to minimize future degradation due to corrosion.

A similarly low value of resistance between widely separated points on a ground reference plane or network ensures that all junctions are well made and that reasonably adequate quantities of conductors are provided throughout the plane or network. In this way, resistive voltage drops are minimized which helps with noise control. In addition, the low value of resistance tends to force the use of reasonably sized conductors which helps minimize path inductance.

It should be recognized that a low dc bond resistance is not a reliable indicator of the performance of the bond at higher frequencies. Inherent conductor inductance and stray capacitance, along with the associated standing wave effects and path resonances, will determine the impedance of the bond. Thus, in rf bonds these factors must be considered along with the dc resistance.

#### 7.4 DIRECT BONDS.

Direct bonding is the establishment of the desired electrical path between the interconnected members without the use of an auxiliary conductor. Specific portions of the surface areas of the members are placed in direct contact. Electrical continuity is obtained by establishing a fused metal bridge across the junction by welding, brazing, or soldering or by maintaining a high pressure contact between the mating surfaces with bolts, rivets, or clamps. Examples of direct bonds are the splices between bus bar sections, the connections between lightning down conductors and the earth electrode subsystem, the mating of equipment front panels to equipment racks, and the mounting of connector shells to equipment panels.

Properly constructed direct bonds exhibit a low dc resistance and provide an rf impedance as low as the configuration of the bond members will permit. Direct bonding is always preferred; however, it can be used only when the two members can be connected together and can remain so without relative movement. The establishment of electrical continuity across joints, seams, hinges, or fixed objects that must be spatially separated requires indirect bonding with straps, jumpers, or other auxiliary conductors.

Current flow through two configurations of a direct bond is illustrated in Figure 7-2. The resistance,  $R_C$ , of the path through the conductors on either side of the bond is given by

$$R_C = \rho \frac{l}{A} \quad (7-1)$$

where  $\rho$  is the resistivity of the conductor materials,  $l$  is the total path length of the current through conductors, and  $A$  is the cross-sectional area of the conductors (assumed equal). Any bond resistance at the junction will increase the total path resistance. Therefore, the objective in bonding is to reduce the bond resistance to a value negligible in comparison to the conductor resistance so that the total path resistance is primarily determined by the resistance of the conductors.

Metal flow processes such as welding, brazing, and silver soldering provide the lowest values of bond resistance. With such processes, the resistance of the joint is determined by the resistivity of the weld or filler metal which can approach that of the metals being joined. The bond members are raised to temperatures sufficient to form a continuous metal bridge across the junction.

For reasons of economy, future accessibility, or functional requirements, metal flow processes are not always the most appropriate bonding techniques. It may then be more appropriate to bring the mating surfaces together under high pressure. Auxiliary fasteners such as bolts, screws, rivets or clamps are employed to apply and maintain the pressure on the surfaces. The resistance of these bonds is determined by the kinds of metals involved, the surface conditions within the bond area, the contact pressure at the surfaces, and the cross-sectional area of the mating surfaces.



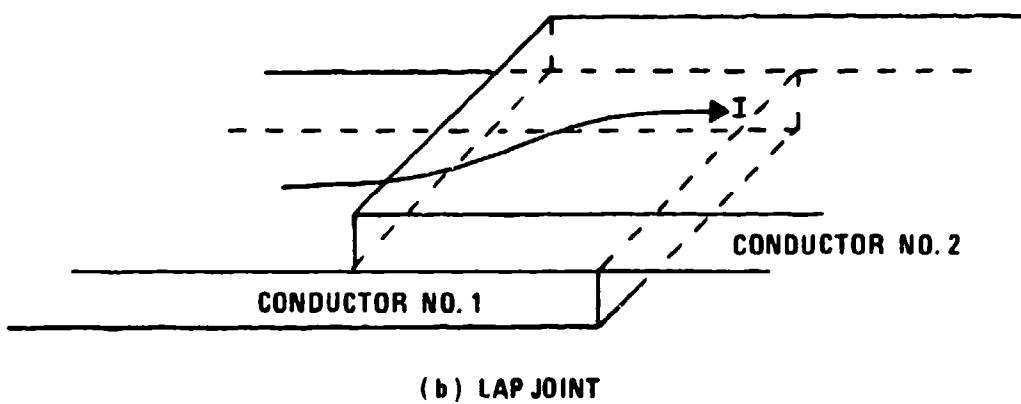
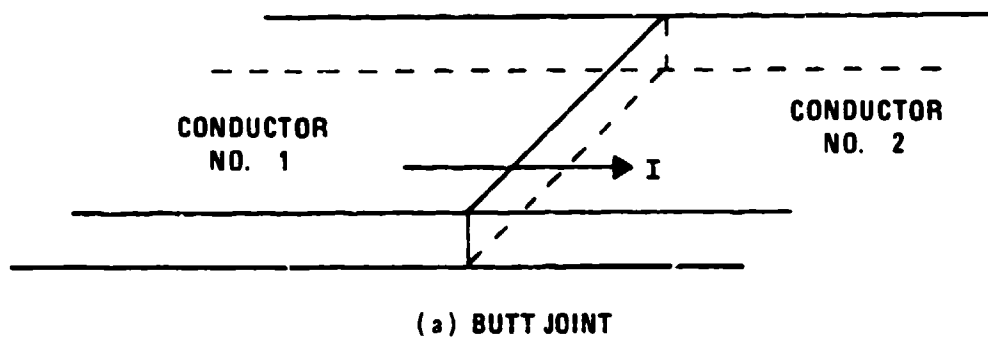


Figure 7-2. Current Flow Through Direct Bonds

#### 7.4.1 Contact Resistance.

No metallic surface is perfectly smooth. In fact, surfaces consist of many peaks and valleys. Even the smoothest commercial surfaces exhibit an RMS roughness of 0.5 to 1 millionth of an inch (7-2); the roughness of most electrical bonding surfaces will be several orders of magnitude greater. When two such surfaces are placed in contact, they touch only at the tips of the peaks - so called asperities. Thus the actual area of contact for current flow is much smaller than the apparent area of metallic contact.

An exaggerated side view of the actual contact surfaces at a bond interface is shown in Figure 7-3. Theoretically, two infinitely hard surfaces would touch at only three asperities. Typically, however, under pressure elastic deformation and plasticity allows other asperities to come into contact. Current passes between the surfaces only at those points where the asperities have been crushed and deformed (7-3) to establish true metal contact. The actual area of electrical contact is equal to the sum of the individual areas of contacting asperities. This actual area of contact can be as little as one millionth of the apparent (gross surface) contact area (7-4).

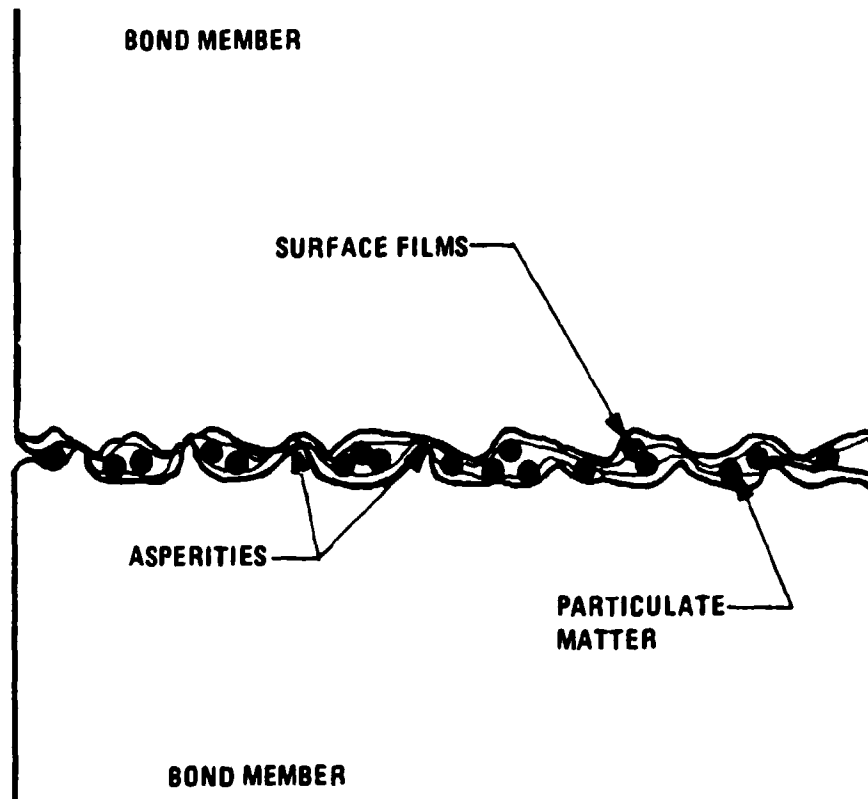


Figure 7-3. Nature of Contact Between Bond Members

#### 7.4.1.1 Surface Contaminants.

Surface films will be present on practically every bond surface. The more active metals such as iron and aluminum readily oxidize to form surface films while the noble metals such as gold, silver, and nickel are less affected by oxide films. Of all metals, gold is the least affected by oxide films. Although silver does not oxidize severely, silver sulfide forms readily in the presence of sulfur compounds.

If the surface films are much softer than the contact material, they can be squeezed from between the asperities to establish a quasi-metallic contact. Harder films, however, may support all or part of the applied load, thus reducing or eliminating the conductive contact area. If such films are present on the bond surfaces, they must be removed through some thermal, mechanical, or chemical means before joining the bond members. Even when metal flow processes are used in bonding, these surface films must be removed or penetrated to permit a homogeneous metal path to be established.

Foreign particulate matter on the bond surfaces will further impair bonding. Dirt and other solid matter such as high resistance metal particles or residue from abrasives can act as stops to prevent metallic contact. Therefore, all such materials must be thoroughly removed from the surfaces prior to joining the bond members.

7.4.1.2 Surface Hardness. The hardness of the bond surfaces also affects the contact resistance. Under a given load, the asperities of softer metals will undergo greater plastic deformation and establish greater metallic contact. Likewise, at a junction between a soft and a hard material, the softer material will tend to conform to the surface contours of the harder material and will provide a lower resistance contact than would be afforded by two hard materials. Table 7-1 shows how the resistance of 6.45 square cm (1 square inch) bonds varies with the type of metals being joined.

7.4.1.3 Contact Pressure. The influence of mechanical load on bond resistance is illustrated by Figure 7-4. This figure shows the resistance variation of a 6.45 square cm (1 square inch) bond held in place with a 1/4-20 steel bolt as a function of the torque applied to the bolt. The resistance variation for brass is lowest due to its relative softness and the absence of insulating oxide films. Even though aluminum is relatively soft, the insulating properties of aluminum oxide cause the bond resistance to be highly dependent upon fastener torque up to approximately 40 in.-lb torque (which corresponds to a contact pressure of about 1200 psi). Steel, being harder and also susceptible to oxide formations, exhibits a resistance that is dependent upon load below 80 in.-lb or about 1500 psi (for mild steel). Above these pressures, no significant improvement in contact resistance can be expected.

Table 7-1

## DC Resistance of Direct Bonds Between Selected Metals

<u>Bond Composition</u>	<u>Resistance (Micro-ohms)</u>
Brass-Brass	6
Aluminum-Aluminum	25
Brass-Aluminum	50
Brass-Steel	150
Aluminum-Steel	300
Steel-Steel	1500

Notes: Apparent Bond Area: 1 in<sup>2</sup> (6.45 cm<sup>2</sup>)  
Fastener Torque: 100 in -lb

Source: Adapted from Reference (7-5)

7.4.1.4 Bond Area.

Smaller bond areas with the same loadings would produce higher contact pressure which would decrease the resistance. However, as shown in Figure 7-4, an increase in pressure over 1500 psi for steel and 1200 psi for aluminum produces relatively slight changes in bond resistance. Further, the improvement in resistance due to increased pressure is offset by the smaller overall bond area. In a similar fashion, a larger bond area (with no change in fastener size) under the same torque results in a lowered pressure at the bond surfaces. The reduced pressure would be counterbalanced to some extent by the increased bond area, but the net effect can be expected to be an increase in bond resistance. Thus, when larger bond areas are used, larger bolts at correspondingly higher torques should be used for fastening. (See Para 7.4.2.4)

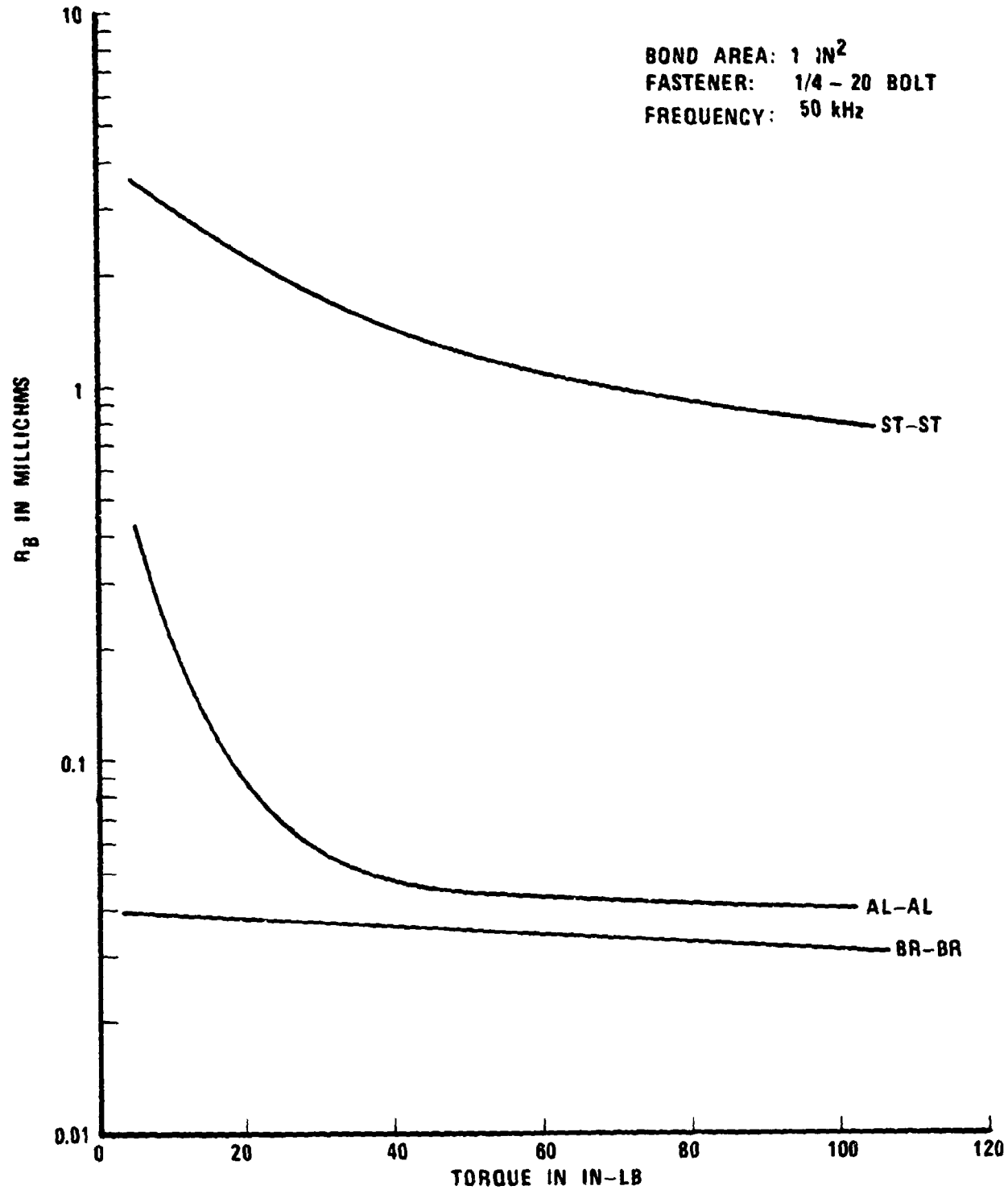


Figure 7-4. Resistance of a Test Bond as a Function of Fastener Torque (7-5)

Bond mating surfaces with areas as large as practical are desirable for several reasons. Large surface areas maximize the cross-sectional area of the path for current and correspondingly maximizes the total number of true metallic contacts between the surfaces. In addition to the obvious advantage of decreased bond resistance, the current crowding which can occur during power fault conditions or under a severe lightning discharge is lessened. Such current crowding produces a higher effective bond resistance than is present during low current flow. The increased bond resistance raises the voltage drop across the junction to even higher values and adds to the heat generated at the junction by the heavy current flow. Large bond areas not only lessen the factors which contribute to heat generation, they also distribute the heat over a larger metallic area which facilitates its removal. A further advantage of a large bond is that it will probably provide greater mechanical strength and will be less susceptible to long term erosion by corrosive products because only a small portion of the total bond area is exposed to the environment.

#### 7.4.2 Direct Bonding Techniques.

Direct bonds may be either permanent or semi-permanent in nature. Permanent bonds may be defined as those intended to remain in place for the expected life of the installation and not required to be disassembled for inspection, maintenance, or system modifications. Joints which are inaccessible by virtue of their location should be permanently bonded and appropriate steps taken to protect the bond against deterioration.

Many bonded junctions must retain the capability of being disconnected without destroying or significantly altering the bonded members. Junctions which should not be permanently bonded include those which may be broken for system modifications, for network noise measurements, for resistance measurements, and for other related reasons. In addition, many joints cannot be permanently bonded for cost reasons. All such connections not permanently joined are defined as semipermanent bonds. Semipermanent bonds include those which utilized bolts, screws, rivets, clamps and other auxiliary devices for fasteners.

##### 7.4.2.1 Welding.

In terms of electrical performance, welding is the ideal method of bonding. The intense heat (in excess of 4000° F) involved is sufficient to boil away contaminating films and foreign substances. A continuous metallic bridge is formed across the joint; the conductivity of this bridge typically approximates that of the bond members. The net resistance of the bond is essentially zero because the bridge is very short relative to the length of the bond members. The mechanical strength of the bond is high; the strength of a welded bond can approach or exceed the strength of the bond members themselves. Since no moisture or contaminants can penetrate the weld, bond corrosion is minimized. The erosion rate of the metallic bridge should be comparable to that of the base members; therefore, the lifetime of the bond should be as great as that of the bond members.

Welds should be utilized whenever practical for permanently joined bonds. Although welding may be a more expensive method of bonding, the reliability of the joint makes it very attractive for bonds which will be inaccessible once construction is completed. Most metals which will be encountered in normal construction can be welded with one of the standard welding techniques such as gas, electric arc, Heliarc and exothermic.

Conventional welding should be performed only by appropriately trained and qualified personnel. Consequently, increased labor costs can be expected. In many instances, also, the welding of bonds can be much slower than the installation of fasteners such as bolts or rivets. In such cases, the added costs of welding may force the use of alternate bonding techniques.

An effective welding technique for many bonding applications is the exothermic process. In this process, a mixture of aluminum, copper oxide, and other powders is held in place around the joint with a graphite mold. The mixture is ignited and the heat generated (in excess of 4000° F) reduces the copper oxide to provide a homogeneous copper blanket around the junction. Because of the high temperatures involved, copper materials can be bonded to steel or iron as well as to other copper materials.

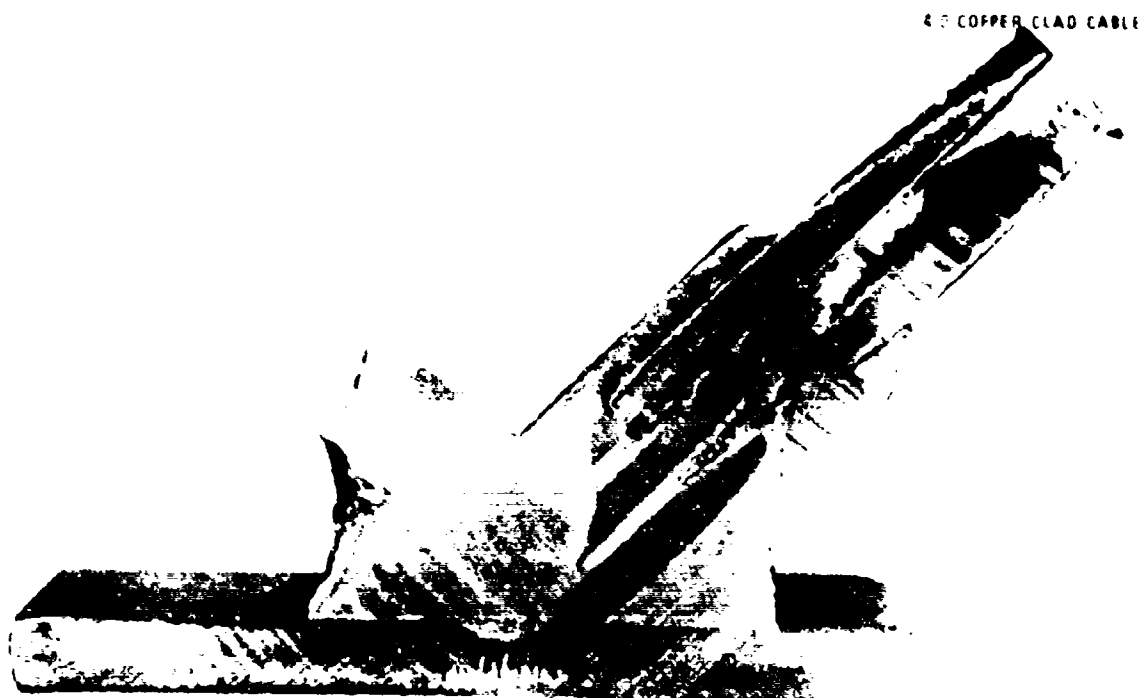
Two examples of exothermic bonds are shown in Figure 7-5. The top photograph shows a 4/0 copper clad cable bonded to a steel plate. The bottom photograph shows two 4/0 copper clad cables axially bonded together. The tight mechanical bond established by this process is evident from these photographs. Figure 7-6 shows examples of the various bond configurations for which molds are readily available.

This process is advantageous for welding cables together, for welding cables to rods, or for welding cables to I-beams and other structural members. It is particularly attractive for the bonding of interconnecting cables to ground rods where the use of conventional welding techniques might be awkward or where experienced welders are not available. Because of the cost of the molds (a separate mold is necessary for each different bond configuration), this process is most economical when there are several bonds of the same configuration to be made.

When using this process, the manufacturer's directions should be followed closely. The mold should be dried or baked out as specified, particularly when the mold has not been used for several hours and may have absorbed moisture. The metals to be bonded should be cleaned of dirt and debris and should have the excess water dried off. Water, dirt and other foreign materials cause voids in the weld which may weaken it or may prevent a low resistance joint from being achieved. A further requirement is that the mold size must match the cable or conductor cross sections; otherwise, the molten metal will not be confined to the bond region.

#### 7.4.2.2 Brazing.

Brazing to include silver soldering is another metal flow process for permanent bonding. In brazing, the bond surfaces are heated to a temperature above 800° F but below the melting point of the bond members. A filler metal with an appropriate flux is applied to the heated members which wets the bond surfaces to provide intimate contact between the brazing solder and the bond surfaces. As with higher temperature welds, the resistance of the brazed joint is essentially zero. However, since brazing frequently involves the use of metal different from the primary bond members, additional precautions must be taken to protect the bond from deterioration through corrosion.



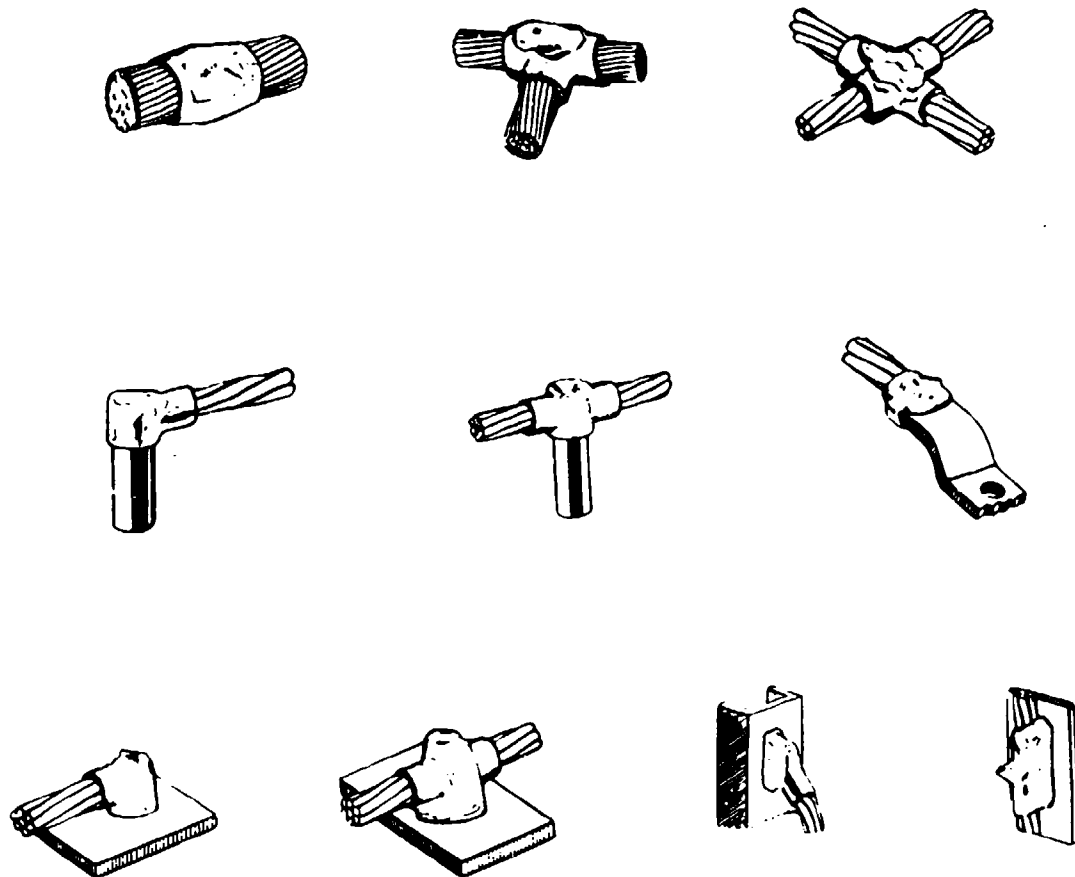
STEEL PLATE

40 COPPER CLAD CABLES



Figure 7-5. Typical Exothermic Connections





COURTESY OF ERICO PRODUCTS, INC., CLEVELAND, OHIO

Figure 7-6. Typical Bond Configurations Which Can Be Implemented With The Exothermic Process

#### 7.4.2.3 Soft Solder.

Soft soldering is an attractive metal flow bonding process because of the ease with which it can be applied. Relatively low temperatures are involved and it can be readily employed with several of the high conductivity metals such as copper, tin and cadmium. With appropriate fluxes, aluminum and other metals can be soldered. Properly applied to compatible materials, the bond provided by solder is nearly as low in resistance as one formed by welding or brazing. Because of its low melting point, however, soft solder should not be used as the primary bonding material where high currents may be present. For this reason, soldered connections are not permitted by MIL-STD-188-124A or the National Electrical Code in grounding circuits for fault protection. Similarly, soft solder is not permitted for interconnections between elements of lightning protection networks by either the Military Standard, the National Fire Protection Association's Lightning Protection Code or the Underwriter's Master Labeled System. In addition to its temperature limitation, soft solder exhibits low mechanical strength and tends to crystallize if the bond members move while the solder is cooling. Therefore, soft solder should not be used if the joint must withstand mechanical loading. The tendency toward crystallization must also be recognized and proper precautions observed when applying soft solder.

Soft solder can be used effectively in a number of ways, however. For example, it can be used to tin surfaces prior to assembly to assist in corrosion control. Soft solder can be used effectively for the bonding of seams in shields and for the joining of circuit components together and to the signal reference subsystem associated with the circuit. Soft solder is often combined with mechanical fasteners in sweated joints. By heating the joint hot enough to melt the solder, a low resistance filler metal is provided which augments the path established by the other fasteners; in addition, the solder provides a barrier to keep moisture and contaminants from reaching the mating surfaces.

#### 7.4.2.4 Bolts.

In many applications, permanent bonds are not desired. For example, equipments must be removed from enclosures or moved to other locations which require that ground leads and other connections must be broken. Often, equipment covers must be removable to facilitate adjustments and repairs. Under such circumstances, a permanently joined connection could be highly inconvenient to break and would limit the operational flexibility of the system. Besides offering greater flexibility, less permanent bonds may be easier to implement, require less operator training, and require less specialized tools.

The most common semipermanent bond is the bolted connection (or one held in place with machine screws, lag bolts, or other threaded fasteners) because this type bond provides the flexibility and accessibility that is frequently required. The bolt (or screw) should serve only as a fastener to provide the necessary force to maintain the 1200-1500 psi pressure required between the contact surfaces for satisfactory bonding. Except for the fact that metals are generally necessary to provide tensile strength, the fastener does not have to be conductive. Although the bolt or screw threads may provide an auxiliary current path through the bond, the primary current path should be established across the metallic interface. Because of the poor reliability of screw thread bonds, self-tapping screws are never to be used for bonding purposes. Likewise, Tinnerman nuts, because of their tendency to vibrate loose, should not be used for securing screws or bolts intended to perform a bonding function.

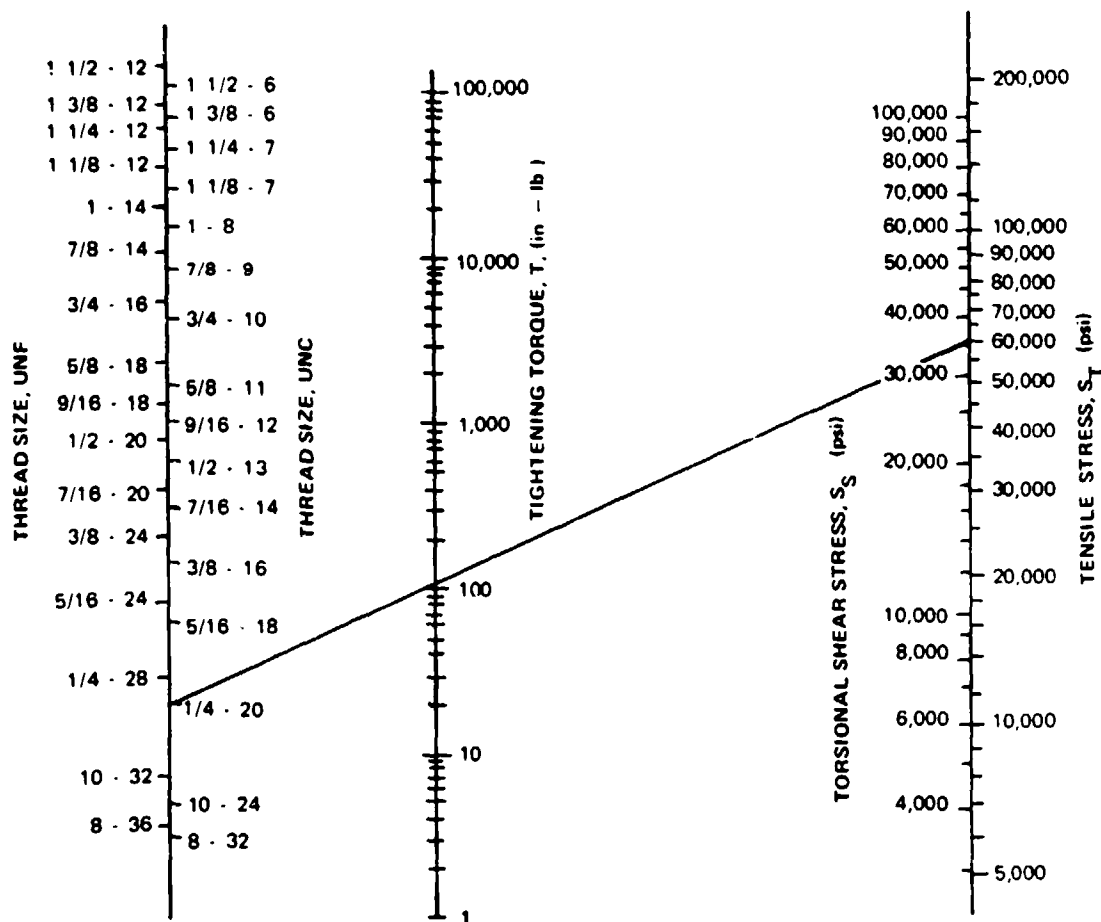


Figure 7-7. Nomograph for Torque on Bolts (7-6)

The size, number and spacing of the fasteners should be sufficient to establish the required bonding pressure over the entire joint area. The pressure exerted by a bolt is concentrated in the immediate vicinity of the bolt head. However, large, stiff washers can be placed under the bolt head to increase the effective contact area. Because the load is distributed over a larger area, the tensile load on the bolt should be raised by increasing the torque. The nomograph of Figure 7-7 may be used to calculate the necessary torque for the size bolts to be used. Where the area of the mating surfaces is so large that unreasonably high bolt torques are required, more than one bolt should be used. For very large mating areas, rigid backing plates should be used to distribute the force of the bolts over the entire area.

#### 7.4.2.5 Rivets.

Riveted bonds are less desirable than bolted connections or joints bridged by metal flow processes. Rivets lack the flexibility of bolts without offering the degree of protection against corrosion of the bond surface that is achieved by welding, brazing or soldering. The chief advantage of rivets is that they can be rapidly and uniformly installed with automatic tools.

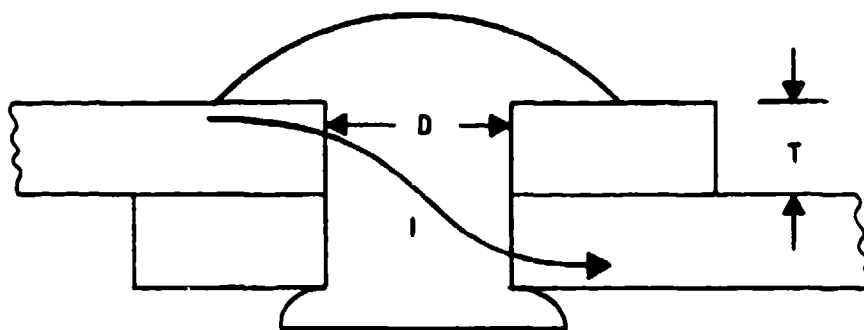
The bonding path established by a rivet is illustrated in Figure 7-8. The current path through a rivet is theorized to be through the interface between the bond members and the rivet body. This theory is justified by experience which shows that the fit between the rivet and the bond members is more important than the state of the mating surfaces between the bond members. Therefore, the hole for the rivet must be a size that provides a close fit to the rivet after installation. The sides of the hole through the bond members must be free of paint, corrosion products, or other non-conducting material.

For riveted joints in shields, the maximum spacing between rivets is recommended to be approximately 2 cm (3/4 inch) or less (7-7). In relatively thin sheet metal, rivets can cause bowing of the stock between the rivets as shown by Figure 7-9. In the bowed or warped regions, metal-to-metal contact may be slight or nonexistent. These open regions allow rf energy to leak through and can be a major cause of poor rf shield performance. By spacing the rivets close together, warping and bowing are minimized. For maximum rf shielding, the seam should be gasketed with some form of wire mesh or conductive epoxy to supplement the bond path of the rivets.

**7.4.2.6 Conductive Adhesive.** Conductive adhesive is a silver-filled, two-component, thermosetting epoxy resin which when cured produces an electrically conductive material. It can be used between mating surfaces to provide low resistance bonds. It offers the advantage of providing a direct bond without the application of heat as is required by metal flow processes. In many locations, the heat necessary for metal flow bonding may pose a fire or explosion threat. When used in conjunction with bolts, conductive adhesive provides an effective metal-like bridge with high corrosion resistance along with high mechanical strength. In its cured state, the resistance of the adhesive may increase through time. It also tends to adhere tightly to the mating surfaces and thus an epoxy-bolt bond is less convenient to disassemble than a simple bolted bond. In some applications, the advantages of conductive adhesive may outweigh this inconvenience.

**7.4.2.7 Comparison of Techniques.** Table 7-2 shows comparative ratings of the most commonly used bonding methods. In this table a rating from zero to 10 is assigned to each method for each performance parameter. A rating of 10 means that the method is suitable from the standpoint of the specific parameter listed in the extreme left hand column of the table. Lower ratings mean that the method is less suitable. A zero rating implies the method is a poor choice, while the dash means it does not apply. One-hundred percent consistency in ratings is impossible because any given method may vary widely in workmanship. A low-rated method expertly performed, will work better than a high-rated poorly performed method. When using the table assume that all methods are equally well implemented.

**7.5 INDIRECT BONDS.** The preferred method of bonding is to connect the objects together with no intervening conductor. Unfortunately, operational requirements or equipment locations often preclude direct bonding. When physical separation is necessary between the elements of an equipment complex or between the complex and its reference plane, auxiliary conductors must be incorporated as bonding straps or jumpers. Such straps are commonly used for the bonding of shock mounted equipment to the structural ground reference. They are also used for by-passing structural elements, such as the hinges on distribution box covers or on equipment covers, to eliminate the wideband noise generated by these elements when illuminated by intense radiated fields or when carrying high level currents. Bond straps or cables are also used to prevent static charge buildup and to connect metal objects to lightning down conductors to prevent flashover.



**D = RIVET DIAMETER**  
**T = THICKNESS OF BOND MEMBER**  
**I = INDICATES CURRENT PATH THROUGH RIVETED BOND**

Figure 7-8. Bonding Path Established by Rivets

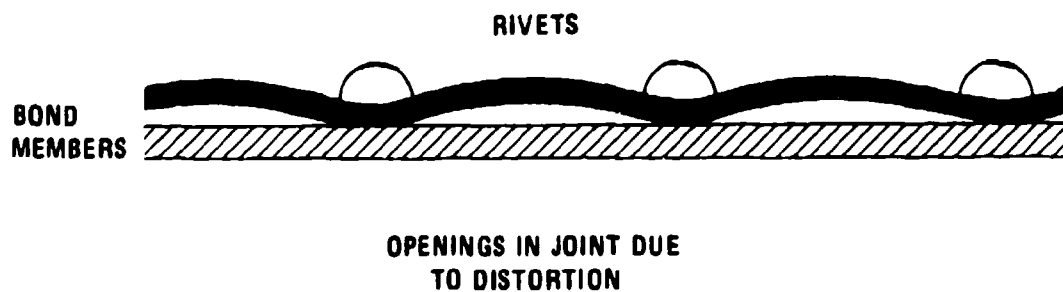


Figure 7-9. An Improperly Riveted Seam

Table 7-2  
 Ratings of Selected Bonding Techniques\*

	Thermal				Chemical			Mechanical					
	Solder	Brazing	Gas or Arc Weld	Exo-thermic Weld	Spot Weld	Con-ductive Adhesive	Bolts	Crimp	Rivets	Clamps	Screws		
											Wire	Terminal	Twisted Wires
Electrical Properties	9	10	10	10	10	3	9	9	9	8	9	9	9
	10	10	10	10	10	7	8	9	9	5	7	9	5
	9	10	10	10	10	5	9	9	10	8	8	9	5
	5	10	10	10	9	1	9	9	9	8	9	9	8
Mechanical Properties	8	10	10	10	10	1	9	9	9	9	3	10	2
	9	10	10	10	10	5	9	9	9	8	3	8	2
	6	10	10	10	7	4	9	9	10	8	5	9	4
	10	10	10	10	10	10	10	10	5	9	10	-	9
Conductor Applicability	9	9	8	9	0	9	9	10	2	9	2	-	9
	5	2	1	8	0	2	9	8	4	9	4	-	1
	8	8	8	9	9	6	9	7	10	7	-	10	-
	3	9	10	10	10	1	8	9	4	5	7	8	5
Environmental	9	10	10	10	10	5	8	9	5	5	8	8	7
	8	9	10	10	10	7	9	8	4	5	2	8	2
	6	7	10	10	9	1	7	9	5	4	2	9	1
	9	10	10	10	10	5	6	9	8	3	7	9	4
Cost Economy	7	4	3	4	4	9	8	8	7	10	9	10	10
	6	5	5	5	5	8	5	9	8	10	10	10	10
	8	5	4	5	8	8	8	9	8	9	9	9	10
	8	6	5	5	9	9	9	9	7	9	10	10	9

Source: Adapted from Reference (7-8)

**7.5.1 Resistance.** The resistance of an indirect bond is equal to the sum of the intrinsic resistance of the bonding conductor and the resistances of the metal-to-metal contacts at each end. The resistance of the strap is determined by the resistivity of the material used and the dimensions of the strap. With typical straps, the dc bond resistance is small. For example with a resistivity of  $1.72 \times 10^{-6}$  ohm-cm, ( $6.77 \times 10^{-7}$  ohm-inches), a copper conductor 2.5 cm, (1 inch) wide, 40 mils thick, and 0.3 meters (1 foot) long has a resistance of 0.2 milliohms. To this resistance will be added the sum of the dc resistances of the direct bonds at the ends of the strap. With aluminum, copper, or brass straps, these resistances should be less than 0.1 milliohm with properly made connections. If long straps are required, however, the resistance of the conductor can be significant (see, for example Table 5-1).

### 7.5.2 Frequency Effects.

**7.5.2.1 Skin Effect.** Because high conductivity materials attenuate radio frequencies rapidly, high frequency currents do not penetrate into conductors very far, i.e., they tend to stay near the surface. At frequencies where this effect becomes significant the ac resistance of the bond strap can differ significantly from its dc value. For a detailed discussion of skin effect, see Section 5.2.2.1.

### 7.5.2.2 Bond Reactance.

The geometrical configuration of the bonding conductor and the physical relationship between objects being bonded introduce reactive components into the impedance of the bond. The strap itself exhibits an inductance that is related to its dimensions. For a straight, flat strap of nonmagnetic metal, the inductance in microhenries is given by

$$L = 0.002\ell \left( 2.303 \log \frac{2\ell}{b+c} + 0.5 + 0.2235 \frac{b+c}{\ell} \right) \mu\text{H} \quad (7-2)$$

or, for a wire of circular cross section, by

$$L = 0.002\ell \left( 2.303 \log \frac{4\ell}{d} - 0.75 \right) \mu\text{H} \quad 7-3$$

where  $\ell$  = length in cm,  
 $b$  = width of the strap in cm  
 $c$  = thickness of the strap in cm, and  
 $d$  = diameter of the wire in cm.

Table 7-3 shows the calculated inductance, using Equation (7-2), of a nonmagnetic rectangular strap, 6 inches (15.2 cm) long. Table 7-4 compares the inductance of 6, 12, and 36 inch lengths of 0.05 inch (1.27 mm) thick straps while Table 7-5 tabulates the inductance of 6, 12, and 36 inch lengths of selected standard size cables from No. 14 AWG to 4/0 AWG. The inductive reactance of the straps tabulated in Tables 7-4 and 7-5 is plotted in Figure 7-10 for frequencies up to 100 MHz.

Table 7-3

Calculated Inductance of a 6 inch (15.2 cm) Rectangular Strap

<u>Width, b</u> (in.)	<u>Thickness, c</u> (in.)	<u>L</u> ( $\mu$ H)
0.5 (12.7 mm)	0.01 (0.25 mm)	0.112
0.5	0.05 (1.27 mm)	0.110
0.5	0.10 (2.54 mm)	0.107
1.0 (25.4 mm)	0.01	0.092
1.0	0.05	0.091
1.0	0.10	0.089
2.0 (50.8 mm)	0.01	0.072
2.0	0.05	0.071
2.0	0.10	0.071

Table 7-4

Calculated Inductance ( $\mu$ H) of 0.05 Inch (1.27 mm) Thick Straps

<u>Width (in.)</u>	<u>Length</u>		
	<u>6 in. (15.2 cm)</u>	<u>12 in. (30.4 cm)</u>	<u>36 in. (91 cm)</u>
0.5 (12.7 mm)	0.110	0.261	0.984
1.0 (25.4 mm)	0.091	0.222	0.866
2.0 (50.8 mm)	0.071	0.183	0.745



Table 7-5

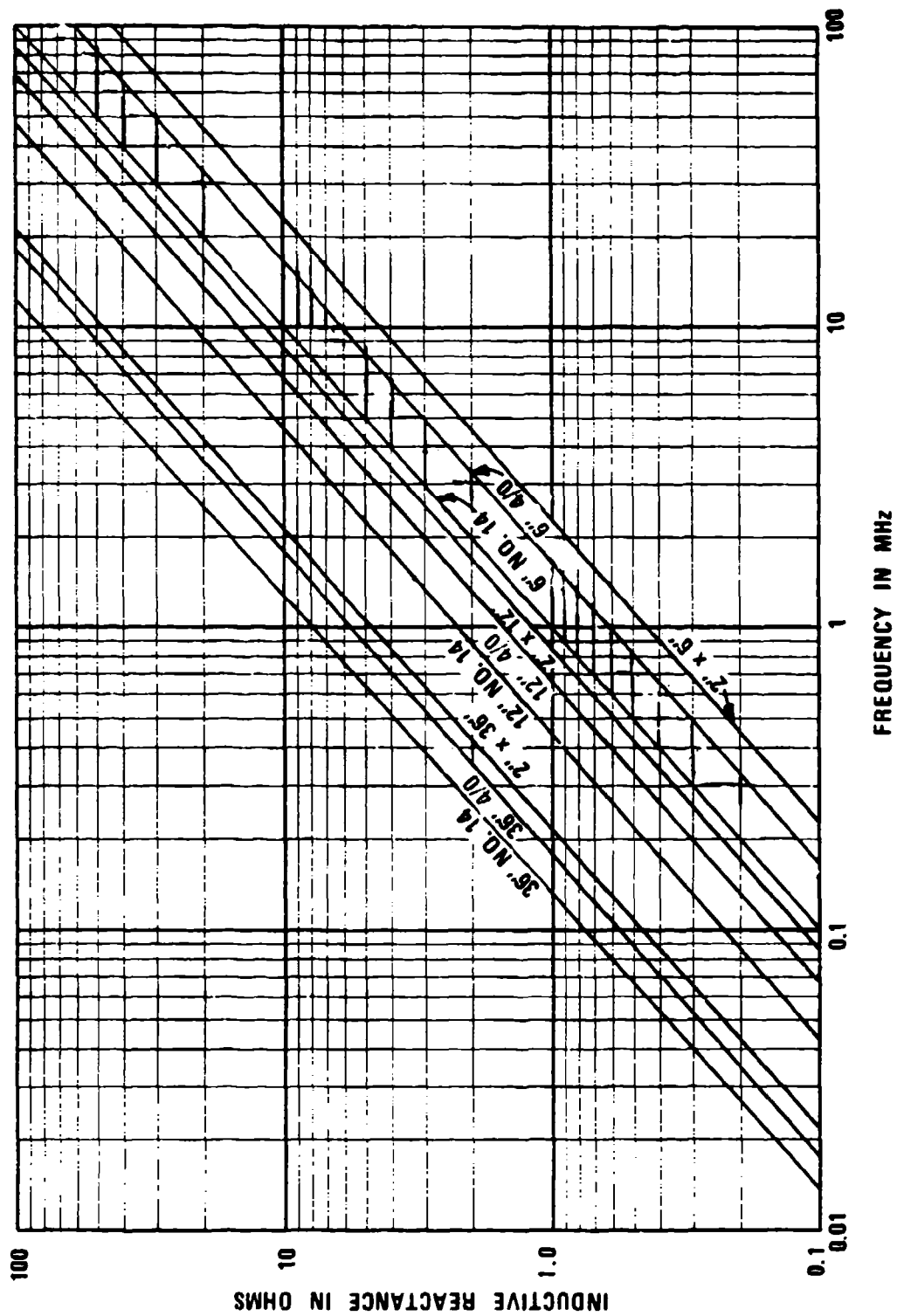
Calculated Inductance ( $\mu\text{H}$ ) of Standard Size Cable

AWG No.	Length		
	6 in.	12 in.	36 in.
4/0	0.098	0.238	0.914
1/0	0.108	0.259	0.977
2	0.115	0.273	1.020
4	0.122	0.287	1.063
6	0.129	0.301	1.105
10	0.144	0.329	1.189
14	0.158	0.358	1.274

Even at relatively low frequencies, the reactance of the inductive component of the bond impedance becomes much larger than the resistance (7-5), (7-9). Thus, in the application of bonding straps, the inductive properties as well as the resistance of the strap must be considered.

The physical size of the bonding strap is important because of its effect on the rf impedance. As the length,  $l$ , of the strap is increased, its impedance increases nonlinearly for a given width; however, as the width,  $b$ , increases, there is a nonlinear decrease in strap impedance. Figure 7-11 shows that the relative reactance of a strap decreases significantly as the length to width ( $l/b$ ) ratio decreases. The curve shows that a strap with an  $l/b$  ratio of 5 to 1 has an inductive reactance that is 45 percent that of a thin wire (i.e., very high ratio of  $l$  to  $b$ ); a 3 to 1 ratio decreases this percentage to 38 percent. Because of this reduction in reactance, bonding straps which are expected to provide a path for rf currents are frequently recommended to maintain a length-to-width ratio of 5 to 1 or less, with a ratio of 3 to 1 preferred.

In many applications, braided straps are preferred over solid straps because they offer greater flexibility. Figure 7-12 compares the measured impedance properties of a braided copper strap with those of a solid copper strap and shows that no significant difference exists between the impedance of the braided or solid strap for frequencies up to 10 MHz. Other tests (7-9) confirm that there is no essential difference in the rf impedance properties of braided and solid straps of the same dimensions and made of the same materials. Because the strands are exposed they are more susceptible to corrosion; braided straps may be undesirable for use in some locations for these reasons. Fine braided straps also are generally not recommended because of higher impedances at the higher frequencies as well as lower current carrying capacities.



**Figure 7-10. Inductive Reactance of Wire and Strap Bond Jumpers**

### 7.5.2.3 Stray Capacitance.

A certain amount of stray capacitance is inherently present between the bonding jumper and the objects being bonded as well as between the bonded objects themselves. Figure 7-13 shows an equivalent circuit for the bonding strap alone.  $R_s$  represents the ac resistance of the strap;  $L_s$  is the inductance which may be calculated with either Equation 7-2 or 7-3; and  $C_s$  is the stray capacitance between the jumper and the two members being bonded. Except for extremely short straps, the magnitude of the inductive reactance of the strap will be significantly larger than the resistance and, at frequencies above approximately 100 kHz, the  $R_s$  term can be ignored. Thus, not considering  $R_s$ , the equation for the magnitude of impedance,  $Z_s$ , of the equivalent circuit is

$$|Z_s| = \frac{\omega L_s}{1 - \omega^2 L_s C_s} \quad (7-4)$$

The equivalent circuit of Figure 7-13 does not take into account the effects of the equipment enclosure or other objects being bonded. Figure 7-14 shows the true equivalent circuit of an indirectly bonded system. The bonding strap parameters are again represented by  $R_s$ ,  $C_s$ ,  $L_s$ .

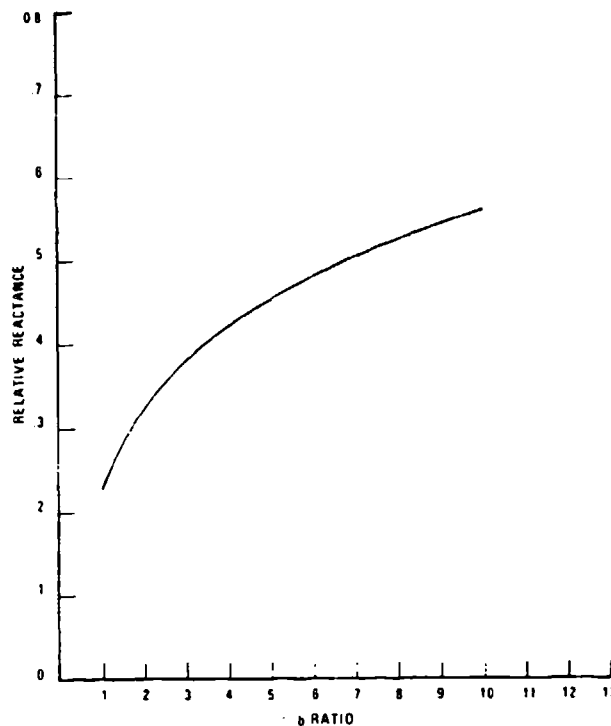


Figure 7-11. Relative Inductive Reactance versus Length-to-Width Ratio of Flat Straps (7-10)

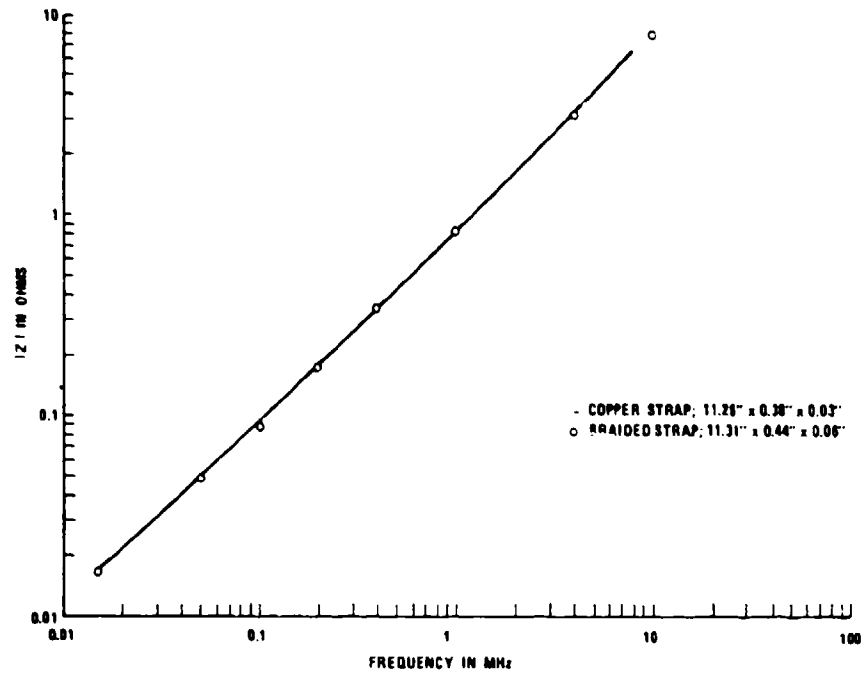


Figure 7-12. Frequency Variation of the Impedance of Simple Conductors (7-5)

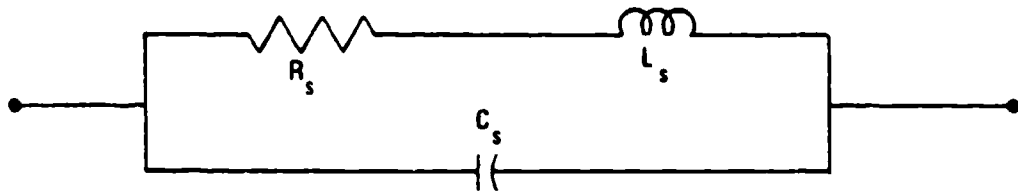


Figure 7-13. Equivalent Circuit for Bonding Strap

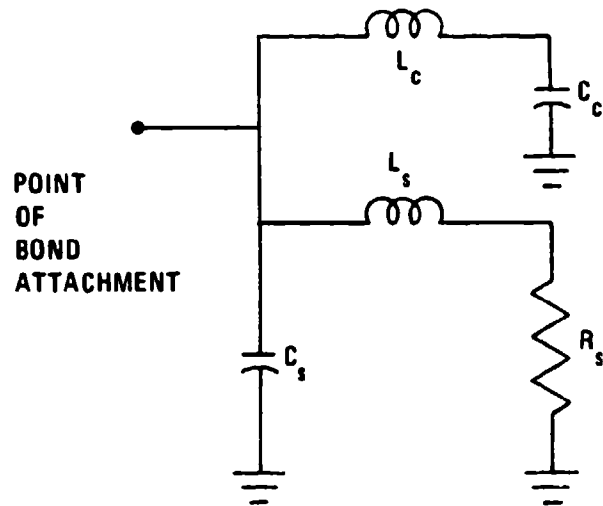


Figure 7-14. True Equivalent Circuit of a Bonded System

The inherent inductance of a bonded object, e.g., an equipment rack or cabinet, is represented by  $L_C$  and the capacitance between the bonded members, i.e., between the equipment and its reference plane, is represented by  $C_C$ . In most situations,  $L_S \gg L_C$ ,  $C_C \gg C_S$ , and  $R_S$  can again be ignored. Thus, the primary (i.e., the lowest) resonant frequency is given by

$$f_r = \frac{1}{2\pi\sqrt{L_S C_C}} \quad (7-5)$$

These resonances can occur at surprisingly low frequencies -- as low as 10 to 15 MHz (7-5) in typical configurations. In the vicinity of these resonances, bonding path impedances of several hundred ohms are common. Because of such high impedances, the strap is not effective. In fact, in these high impedance regions, the bonded system may act as an effective antenna system which increases the pickup of the same signals which bond straps are intended to reduce. Figures 7-15 and 7-16 show the measured effectiveness of two different lengths of bonding straps in the reduction of the voltage induced by a radiated field on an equipment cabinet above a ground plane. The bond effectiveness indicates the amount of voltage reduction achieved by the addition of the bonding strap. Positive values of bonding effectiveness indicate a lowering of the induced voltage. At frequencies near the network resonances, the induced voltages are higher with the bonding straps than without the straps. Figures 7-15 and 7-16 show that:

- a. at low frequencies where the reactance of the strap is low, bonding straps will provide effective bonding;
- b. at frequencies where parallel resonances exist in the bonding network, straps may severely enhance the pickup of unwanted signals; and
- c. above the parallel resonant frequency, bonding straps do not contribute to the pickup of radiated signals either positively or negatively.

In conclusion, bonding straps should be designed and used with care with special note taken to ensure that unexpected interference conditions are not generated by the use of such straps.

**7.6 SURFACE PREPARATION.** To achieve an effective and reliable bond, the mating surfaces must be free of any foreign materials, e.g., dirt, filings, preservatives, etc., and nonconducting films such as paint, anodizing, and oxides and other metallic films. Various mechanical and chemical means can be used to remove the different substances which may be present on the bond surfaces. After cleaning, the bond should be assembled or joined as soon as possible to minimize recontamination of the surfaces. After completion of the joining process, the bond region should be sealed with appropriate protective agents to prevent bond deterioration through corrosion of the mating surfaces.

### 7.6.1 Solid Materials.

Solid material such as dust, dirt, filings, lint, sawdust and packing materials impede metallic contact by providing mechanical stops between the surfaces. They can affect the reliability of the connection by fostering corrosion. Dust, dirt, and lint will absorb moisture and will tend to retain it on the surface. They may even promote the growth of molds, fungi, and bacteriological organisms which give off corrosive products. Filings of foreign metals can establish tiny electrolytic cells (see Section 7.8) which will greatly accelerate the deterioration of the surfaces.

The bond surface should be cleaned of all such solid materials. Mechanical means such as brushing or wiping are generally sufficient. Care should be exercised to see that all materials in grooves or crevices are removed. If a source of compressed air is available, air blasting is an effective technique for removing solid particles if they are dry enough to be dislodged.

### 7.6.2 Organic Compounds.

Paints, varnishes, lacquers, and other protective compounds along with oils, greases and other lubricants are nonconductive and, in general, should be removed. Commercial paint removers can be used effectively. Lacquer thinner works well with oil-based paints, varnish, and lacquer. If chemical solvents cannot be used effectively, mechanical removal with scrapers, wire brushes, power sanders, sandpaper, or blasters should be employed. When using mechanical techniques, care should be exercised to avoid removing excess material from the surfaces. Final cleaning should be done with a fine, such as 400-grit, sandpaper or steel wool. After all of the organic material is removed, abrasive grit or steel wool filaments should be brushed or blown away. A final wipe down with denatured alcohol, dry cleaning fluid, or lacquer thinner should be accomplished to remove any remaining oil or moisture films.

### WARNING

Many paint solvents such as lacquer thinner and acetone are highly flammable and toxic in nature. They should never be used around open flames and adequate ventilation must be present. Inhalation of the fumes must be prevented.

Oils, greases, and other petroleum compounds should be wiped with a cloth or scraped off. Residual films should be dissolved away with an appropriate solvent. Hot soapy water can be used effectively for removing any remaining oil or grease. If water is used, however, the surfaces must be thoroughly dried before completing the bond. For small or intricate parts, vapor degreasing is an effective cleaning method. Parts to be cleaned are exposed to vapors of trichlorethylene, perchlorethylene, or methylene chloride until the surfaces reach the temperature of the vapor. In extreme cases, further cleaning by agitation in a bath of dry chromic acid, 2 lbs per gallon of water, and sulfuric acid, 4 oz per gallon of water, (7-7) may be necessary. The average dip time should be restricted to less than 30 seconds because prolonged immersion of parts in this bath may produce severe etching and cause loss of dimension. This bath must be followed by a thorough rinse with cold water and then a hot water rinse to facilitate drying.

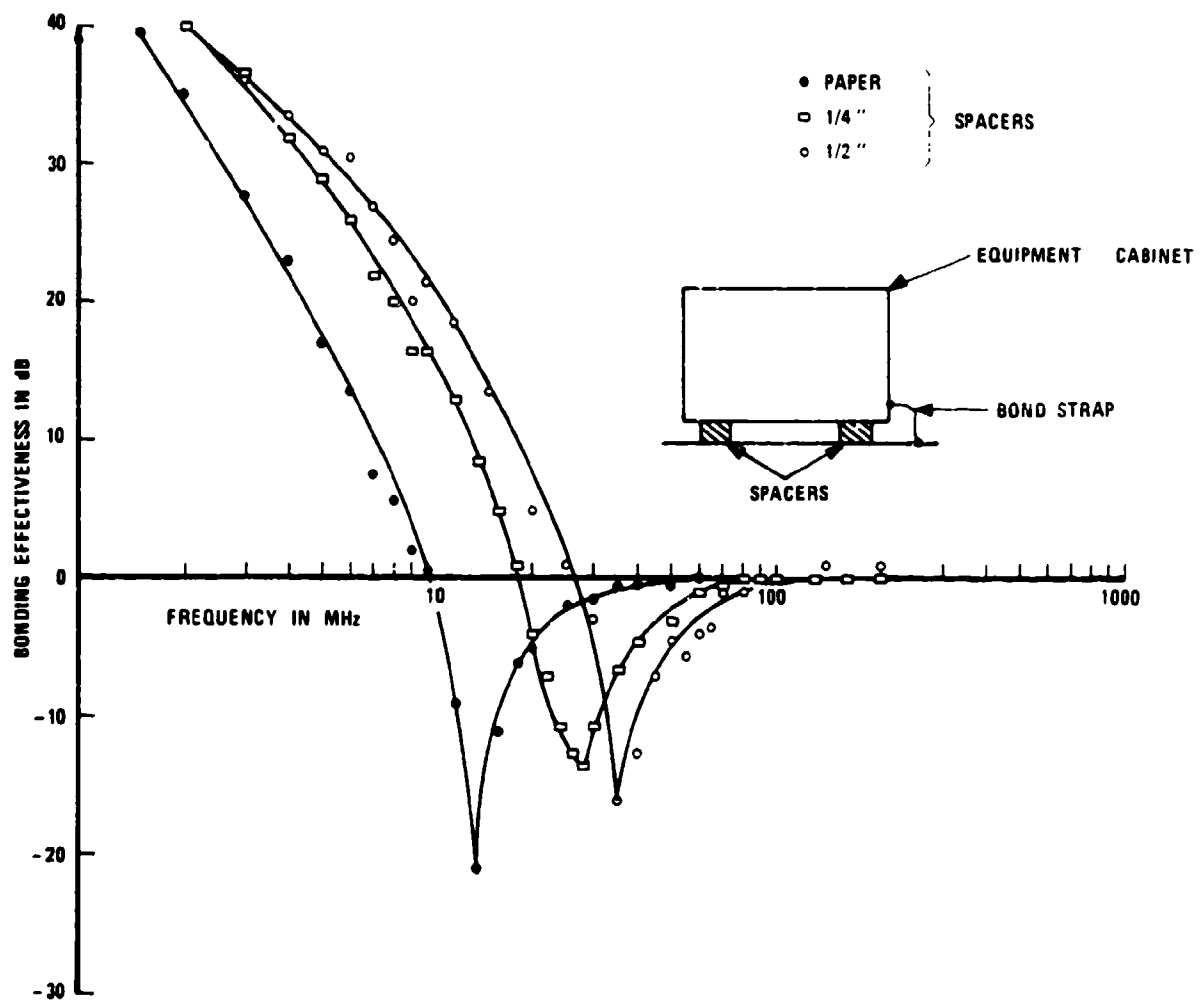


Figure 7-15. Measured Bonding Effectiveness of a 9-1/2 Inch Bonding Strap (7-5)

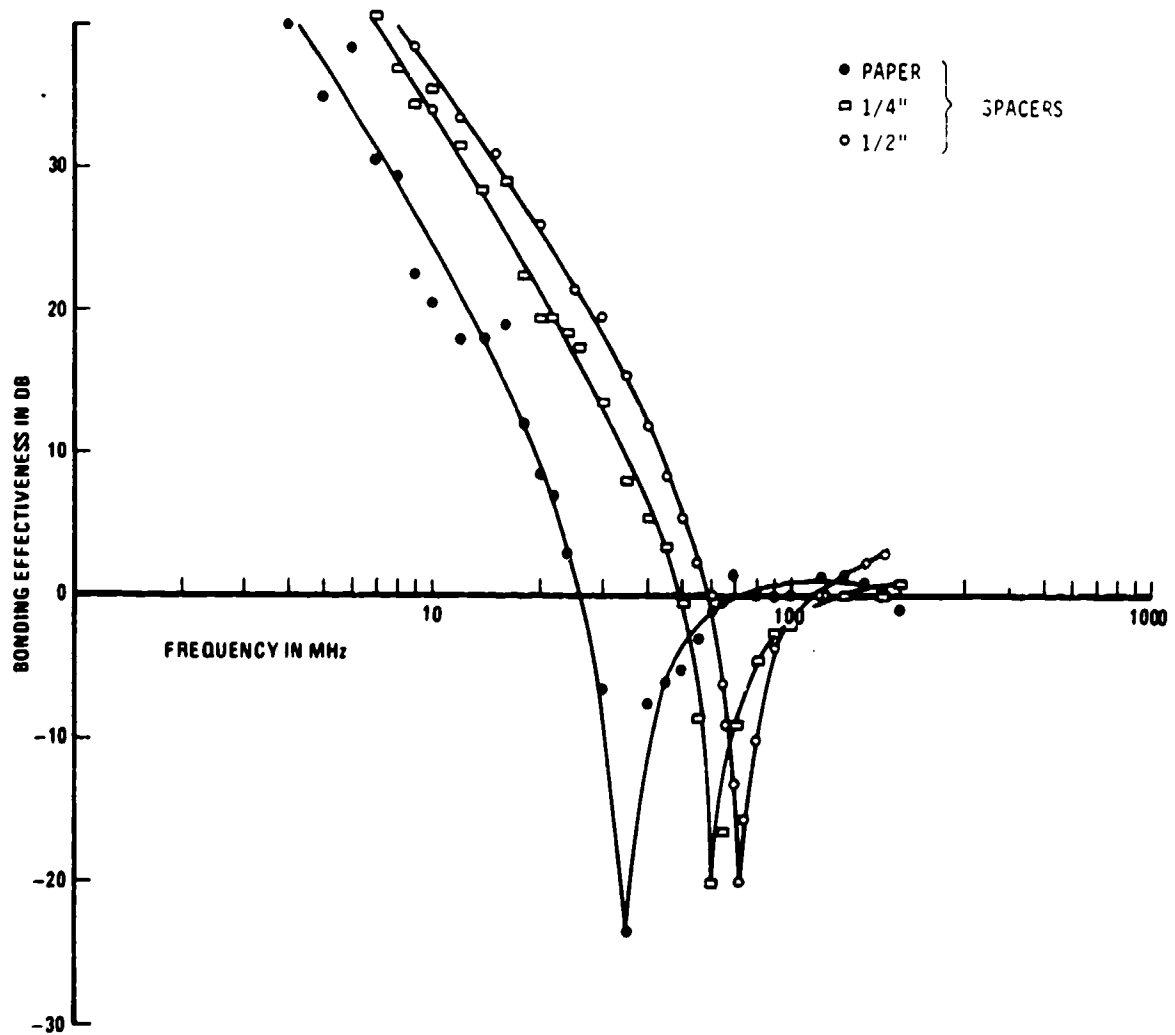


Figure 7-16. Measured Bonding Effectiveness of a 2-3/8 Inch Bonding Strap (7-5)



### 7.6.3 Platings and Inorganic Finishes.

Many metals are plated or coated with other metals or are treated to produce surface films to achieve improved wearability or provide corrosion resistance. Metal platings such as gold, silver, nickel, cadmium, tin, and rhodium should have all foreign solid materials removed by brushing or scraping and all organic materials removed with an appropriate solvent. Since such platings are usually very thin, acids and other strong etchants should not be used. Once the foreign substances are removed, the bond surfaces should be burnished to a bright shiny condition with fine steel wool or fine grit sandpaper. Care must be exercised to see that excessive metal is not removed. Finally, the surfaces should be wiped with a cloth dampened in a denatured alcohol or dry cleaning solvent and allowed to dry before completing the bond.

Chromate coatings such as Iridite 14, Iridite 18P, Oadkrite 36, and Alodine 1000 offer low resistance as well as provide corrosion resistance. These coatings should not be removed. In general, any chromate coatings meeting the requirements of MIL-C-5541 (7-11) should be left in place.

Many aluminum products are anodized for appearance and corrosion resistance. Since these anodic films are excellent insulators, they must be removed prior to bonding. Those aluminum parts to be electrically bonded either should not be anodized or the anodic coating must be removed from the bond area.

7.6.4 Corrosion By-Products. Oxides, sulfides, sulfates, and other corrosion by-products must be removed because they restrict or prevent metallic contact. Soft products such as iron oxide and copper sulfate can be removed with a stiff wire brush, steel wool, or other abrasives. Removal down to a bright metal finish is generally adequate. When pitting has occurred, refinishing of the surface by grinding or milling may be necessary to achieve a smooth, even contact surface. Some sulfides are difficult to remove mechanically and chemical cleaning and polishing may be necessary. Oxides of aluminum are clear and thus the appearance of the surface cannot be relied upon as an indication of the need for cleaning. Although the oxides are hard, they are brittle and roughening of the surface with a file or coarse abrasive is an effective way to prepare aluminum surfaces for bonding.

### 7.7 COMPLETION OF THE BOND.

After cleaning of the mating surfaces, the bond members should be assembled or attached as soon as possible. Assembly should be completed within 30 minutes if at all possible. If more than 2 hours is required between cleaning and assembly, a temporary protective coating must be applied. Of course, this coating must also be removed before completing the bond.

The bond surfaces must be kept free of moisture before assembly and the completed bond must be sealed against the entrance of moisture into the mating region. Acceptable sealants are paint, silicone rubber, grease, and polysulfates. Where paint has been removed prior to bonding, the completed bond should be repainted to match the original finish. Excessively thinned paint should be avoided; otherwise, the paint may seep under the edges of the bonded components and impair the quality of the connection. Compression bonds between copper conductors or between compatible aluminum alloys located in readily accessible areas not subject to weather exposure, corrosive fumes, or excessive dust do not require sealing. This is subject to the approval of the responsible civil engineer or the local authorized approval representative.

**7.8 BOND CORROSION.** Corrosion is the deterioration of a substance (usually a metal) because of a reaction with its environment. Most environments are corrosive to some degree. Those containing salt sprays and industrial contaminants are particularly destructive. Bonds exposed to these and other environments must be protected to prevent deterioration of the bonding surfaces to the point where the required low resistance connection is destroyed.

**7.8.1 Chemical Basis of Corrosion.**

The basic diagram of the corrosion process for metals is shown in Figure 7-17. The requirements for this process to take place are that (1) an anode and a cathode must be present to form an electrochemical cell and (2) a complete path for the flow of direct current must exist. These conditions occur readily in many environments. On the surface of a single piece of metal, anodic and cathodic regions are present because of impurities, grain boundaries and grain orientations, or localized stresses. These anodic and cathodic regions are in electrical contact through the body of metal. The presence of an electrolyte or conducting fluid completes the circuit and allows the current to flow from the anode to the cathode of the cell.

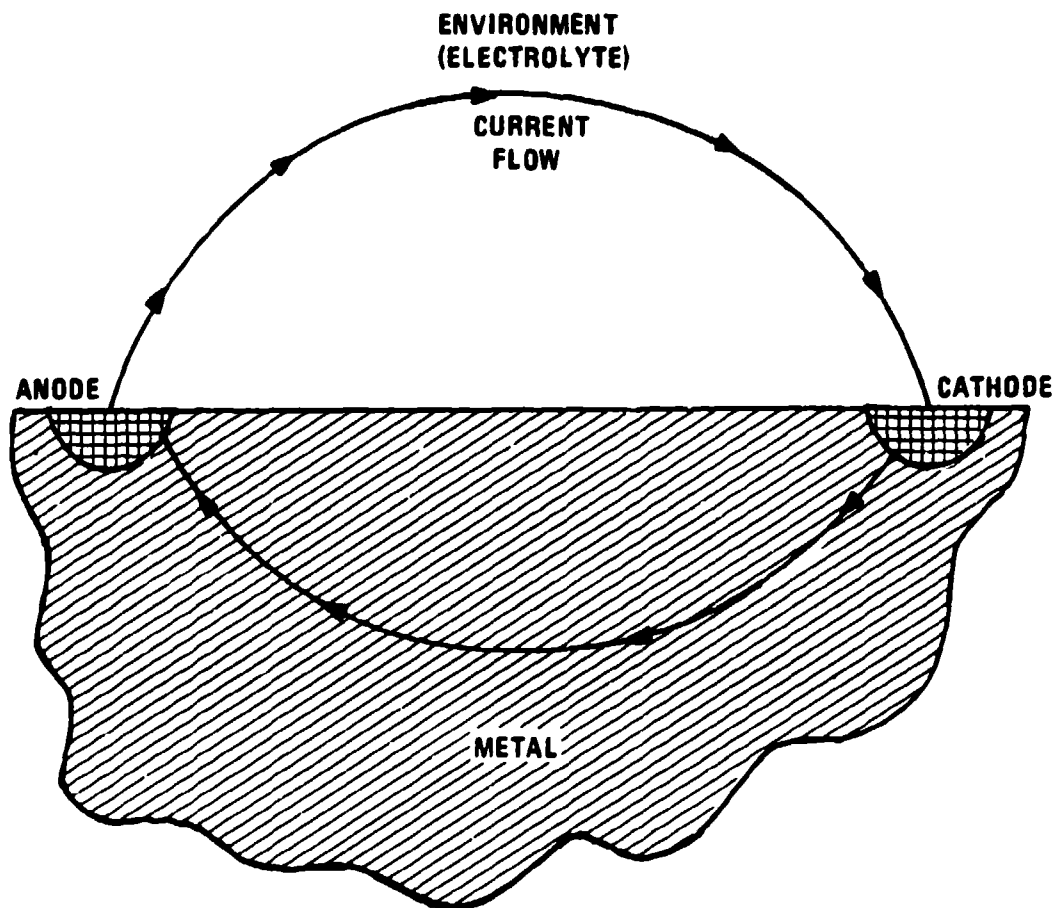


Figure 7-17. Basic Diagram of the Corrosion Process

Anything that prevents the existence of either of the above conditions will prevent corrosion. For example, in pure water, hydrogen gas will accumulate on the cathode to provide an insulating blanket to stop current flow. Most water, however, contains dissolved oxygen which combines with the hydrogen to form additional molecules of water. The removal of the hydrogen permits corrosion to proceed. This principle of insulation is employed in the use of paint as a corrosion preventive. Paint prevents moisture from reaching the metal and thus prevents the necessary electrolytic path from being established.

**7.8.1.1 Electrochemical Series.** The oxidation of metal involves the transfer of electrons from the metal to the oxidizing agent. In this process of oxidation, an electromotive force (EMF) is established between the metal and the solution containing the oxidizing agent. A metal in contact with an oxidizing solution containing its own metal ions establishes a fixed potential difference with respect to every other metal in the same condition. The set of potentials determined under a standardized set of conditions, including temperature and ion concentration in the solution, is known as the EMF (or electrochemical) series. The EMF series (with hydrogen as the referenced potential of 0 volts) for the more common metals is given in Table 7-6. The importance of the EMF series is that it shows the relative tendencies of metals to corrode. Metals high in the series react more readily and are thus more prone to corrosion. The series also indicates the magnitude of the potential established when two metals are coupled to form a cell. The farther apart the metals are in the series, the higher the voltage between them. The metal higher in the series will act as the anode and the one lower will act as the cathode. When the two metals are in contact, loss of metal at the anode will occur through oxidation to supply the electrons to support current flow. This type of corrosion is defined as galvanic corrosion. The greater the potential difference of the cell, i.e., the greater the dissimilarity of the metals, the greater the rate of corrosion of the anode.

**7.8.1.2 Galvanic Series.**

The EMF series is based on metals in their pure state -- free of oxides and other films -- in contact with a standardized solution. Of greater interest in practice, however, is the relative ranking of metals in a typical environment with the effects of surface films included. This ranking is referred to as the galvanic series. The most commonly referenced galvanic series is listed in Table 7-7. This series is based on tests performed in sea water and should be used only as an indicator where other environments are of concern.

Galvanic corrosion in the atmosphere is dependent largely on the type and amount of moisture present. For example, corrosion will be more severe near the seashore and in polluted industrial environments than in dry rural settings. Condensate near the seashore or in industrial environments is more conductive even under equal humidity and temperature conditions due to increased concentration of sulfur and chlorine compounds. The higher conductivity means that the rate of corrosion is increased.

Table 7-6

## Standard Electromotive Series (7-12)

<u>Metal</u>	<u>Electrode Potential*</u> (volts)
Magnesium	2.37
Aluminum	1.66
Zinc	0.763
Iron	0.440
Cadmium	0.403
Nickel	0.250
Tin	0.136
Lead	0.126
Copper	-0.337
Silver	-0.799
Palladium	-0.987
Gold	-1.50

NOTE: \*Signs of potential are those employed by the American Chemical Society.

Table 7-7

Galvanic Series of Common Metals and Alloys in Seawater (7-13)

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---

(ANODIC OR ACTIVE END)

Magnesium  
Magnesium Alloys  
Zinc  
Galvanized Steel or Iron  
1100 Aluminum  
Cadmium  
2024 Aluminum  
Mild Steel or Wrought Iron  
Cast Iron  
Chromium Steel (active)  
Ni-Resist (high-Ni cast iron)  
18-8 Stainless Steel (active)  
18-8 Mo Stainless Steel (active)  
Lead-tin Solders  
Lead  
Tin  
Nickel (active)  
Inconel (active)  
Hastelloy B  
Manganese Bronze  
Brasses  
Aluminum Bronze  
Copper  
Silicon Bronze  
Monel  
Silver Solder  
Nickel  
Inconel  
Chromium Steel  
18-8 Stainless Steel  
18-8 Mo Stainless Steel  
Hastelloy C  
Chlorimet 3  
Silver  
Titanium  
Graphite  
Gold  
Platinum

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(CATHODIC OR MOST NOBLE END)

**7.8.2 Relative Area of Anodic Member.** When joints between dissimilar metals are unavoidable, the anodic member of the pair should be the largest of the two. For a given current flow in a galvanic cell, the current density is greater for a small electrode than for a larger one. The greater the current density of the current leaving an anode, the greater is the rate of corrosion as illustrated by Figure 7-18. As an example, if a copper strap or cable is bonded to a steel column, the rate of corrosion of the steel will be low because of the large anodic area. On the other hand, a steel strap or bolt fastener in contact with a copper plate will corrode rapidly because of the relatively small area of the anode of the cell.

**7.8.3 Protective Coatings.** Paint or metallic platings used for the purpose of excluding moisture or to provide a third metal compatible with both bond members should be applied with caution. When they are used, both members must be covered as illustrated in Figure 7-19. Covering the anode alone must be avoided. If only the anode is covered then at imperfections and breaks in the coating, corrosion will be severe because of the relatively small anode area. All such coatings must be maintained in good condition.

## **7.9 WORKMANSHIP.**

Whichever bonding method is determined to be the best for a given situation, the mating surfaces must be cleaned of all foreign material and substances which would preclude the establishment of a low resistance connection. Next, the bond members must be carefully joined employing techniques appropriate to the specific method of bonding. Finally the joint must be finished with a protective coating to ensure continued integrity of the bond. The quality of the junction depends upon the thoroughness and care with which these three steps are performed. In other words, the effectiveness of the bond is influenced greatly by the skill and conscientiousness of the individual making the connection. Therefore, this individual must be aware of the importance of electrical bonds and must have the necessary expertise to correctly implement the method of bonding chosen for the job.

Those individuals charged with making bonds must be carefully trained in the techniques and procedures required. Where bonds are to be welded, for example, work should be performed only by qualified welders. No additional training should be necessary because standard welding techniques appropriate for construction purposes are generally sufficient for establishing electrical bonds. Qualified welders should also be used where brazed connections are to be made.

Exothermic welding can be effectively accomplished by personnel not specifically trained as welders. Every individual doing exothermic welding should become familiar with the procedural details and with the precautions required with these processes. Contact the manufacturers of the materials for such processes for assistance in their use. By taking reasonable care to see that the bond areas are clean and free of water and that the molds are dry and properly positioned, reliable low resistance connections can be readily achieved.

Pressure bonds utilizing bolts, screws, or clamps must be given special attention. Usual construction practices do not require the surface preparation and bolt tightening necessary for an effective and reliable electrical bond. Therefore, emphasis beyond what would be required for strictly mechanical strength is necessary. Bonds of this type must be checked rigorously to see that the mating surfaces are carefully cleaned, that the bond members are properly joined, and that the completed bond is adequately protected against corrosion.

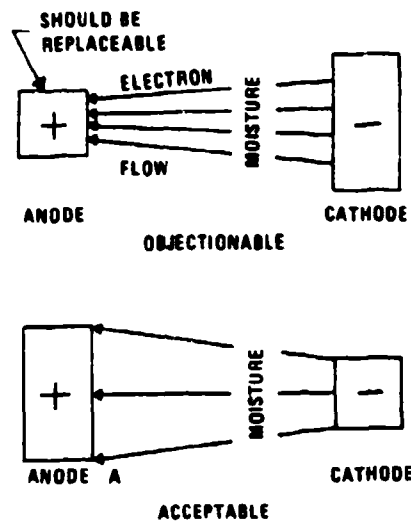


Figure 7-18. Anode-to-Cathode Size at Dissimilar Junctions

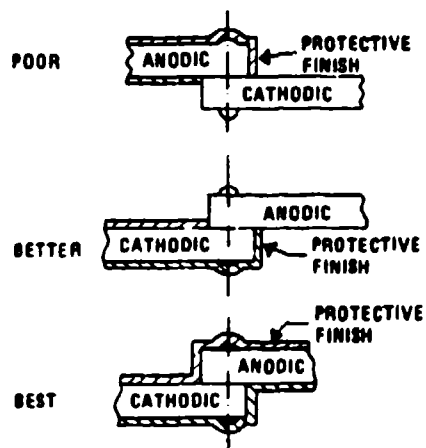


Figure 7-19. Techniques for Protecting Bonds Between Dissimilar Metals

**7.10 SUMMARY OF GUIDELINES.**

- Bonds must be designed into the system. Specific attention should be directed to the interconnections not only in power lines and signal lines, but also between conductors of signal ground bus networks, between equipments and the ground bus networks, between both cable and component or compartment shields and the ground reference plane, between structural members, and between elements of the lightning protection network. In the design and construction of a facility, signal path, personnel safety, and lightning protection bonding requirements must be considered along with mechanical and operational needs.
- Bonding must achieve and maintain intimate contact between metal surfaces. The surfaces must be smooth and clean and free of nonconductive finishes. Fasteners must exert sufficient pressure to hold the surfaces in contact in the presence of the deforming stresses, shocks, and vibrations associated with the equipment and its environment.
- The effectiveness of the bond depends upon its construction, the frequency and magnitude of the currents flowing through it, and the environmental conditions to which it is subjected.
- Bonding jumpers are only a substitute for direct bonds. If the jumpers are kept as short as possible, have a low resistance and low  $L/W$  ratio, and are not higher in the electrochemical series than the bonded members, they can be considered a reasonable substitute.
- Bonds are always best made by joining similar metals. If this is not possible, special attention must be paid to the control of bond corrosion through the choice of the materials to be bonded, the selection of supplementary components (such as washers) to assure that corrosion will affect replaceable elements only, and the use of protective finishes.
- Protection of the bond from moisture and other corrosive elements must be provided.
- Finally, throughout the lifetime of the equipment, system, or facility, the bonds must be inspected, tested, and maintained to assure that they continue to perform as required.



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## CHAPTER 8

## SHIELDING

**8.1 FUNCTION OF AN ELECTROMAGNETIC SHIELD.**

Groups of equipment or subsystems may be made electromagnetically compatible by any combination of three fundamental approaches: (1) the interfering signal source level may be reduced, (2) the receptor susceptibility may be reduced, or (3) the attenuation of the path or paths over which interference is transmitted from source to receptor may be increased. Radiated interference signals generated by electromagnetic fields may be attenuated effectively by electromagnetic shielding, either at the source or at the receptor. An electromagnetic (EM) shield reduces the strength of electric and/or magnetic fields on the side of the shield away from an interfering EM source. When a shield encloses an EM source, the field strength outside the shield will be reduced; when the shield is used to enclose a sensitive (susceptible) assembly located near an external EM source, the field strength inside the enclosure is substantially reduced. Shielding, when properly designed and implemented, offers significant wideband protection against EM radiation where source and receptor are not sufficiently separated for adequate free space radiation attenuation. It is relatively easy to obtain 40 dB of shielding effectiveness in a frequency range above 100 kHz with a single shield, and values as high as 70 dB are readily obtained with careful single-shield construction. For higher values of shielding effectiveness, double shields are normally used, yielding shielding values as high as 120 dB.

Radiated energy may still be coupled into a susceptible device through a shield of inadequate thickness, through holes provided for ventilation and other purposes, and through imperfectly joined shield sections. Precise calculation of shielding effectiveness, even for perfectly joined solid shields, depends on the form of the shield and the type field for which the shielding is to be used. Both electric and magnetic coupling can occur, but normally it is relatively easy to provide electric shielding. Magnetic shielding, however, is more difficult to provide, particularly at frequencies below 100 kHz. To avoid uncertainties in critical situations, tests should be performed to check shielding effectiveness. Such tests require the establishment of a known field and the measurement of insertion loss introduced by the shielding.

In the construction of a facility, the installation designer should take advantage of all the inherent shielding which the installation and its individual equipments and terrain have to offer. Items such as building walls, partitions, towers and other similar structures may be used to advantage. The shielding effectiveness afforded by these items can be used to isolate EM radiation generating equipment from potentially susceptible devices, personnel, flammable mixtures, and other items. In addition, equipments used in a console or rack may be placed to take advantage of the inherent shielding of that rack.

Shielding, although an important technique for reducing EM interference effects, is not the only technique available for this purpose. Application of shielding techniques should not be made without due regard to the roles which filtering, grounding, and bonding play in the interference suppression program.

## 8.2 BASIC SHIELDING THEORY.

The shielding effectiveness of an equipment or subassembly enclosure depends upon a number of parameters, the most notable of which are the frequency and impedance of the impinging wave, the intrinsic characteristics of the shield materials, and the numbers and shapes of shield discontinuities. The effectiveness of a shield is specified in terms of the reduction of EM field strength caused by the shield. The shielding effectiveness (SE) is defined as the ratio of the field strength without the shield present to the field strength with the shield in place. Because of the wide ranges in this ratio, it is common practice to express the shielding effectiveness in decibels:

$$SE = 20 \log (E_1/E_2), \quad (8-1)$$

or

$$SE = 20 \log (H_1/H_2). \quad (8-2)$$

The variables  $E_1$  and  $H_1$  are the electric and magnetic field strengths without the shield present, and  $E_2$  and  $H_2$  are those with the shield in place.

**8.2.1 Oppositely Induced Fields.** A shielding action occurs whenever an electromagnetic wave encounters a metal surface. Part of the wave energy is reflected back toward the source, part is dissipated in the metal, and the remainder propagates beyond the metal. This shielding effect can be visualized as being the result of the incoming electric and magnetic fields inducing charges at the surface of the shield and a current flow within the shield, respectively. The induced charges and currents are of such a polarity and direction that their associated electric and magnetic fields oppose the incident fields, thus reducing the EM fields beyond the shield. Although this concept of the shielding theory does not lend itself to efficient calculation of the degree of shielding provided by a particular shield, it does provide a useful physical picture of shielding. For example, it can be seen from this viewpoint that shielding effectiveness would be reduced more if the shield were cut so as to interfere with the induced current flow than if it were cut along the line of current flow. Thus, if a plane EM wave is incident upon a conducting shield with a very long slit, more energy will be transmitted through the slit if the electric field vector is perpendicular to the slit than if it is parallel to the slit (see Figure 8-1). This is true because the EM boundary conditions (8-1) require that the induced shield current flow be perpendicular to the incoming magnetic field vector (and thus parallel to the electric field vector).

**8.2.2 Transmission Line Analogy.** The shielding theory most applicable to engineering calculations is based upon an analogy with transmission line theory. According to the planewave theory developed by Schelkunoff (8-1), an electromagnetic shield transmits EM waves whose fronts coincide with the shielding boundary configuration in a manner mathematically analogous to that in which a two-wire transmission line transmits electrical current and voltage. Consider an incident EM wave with a power of  $P_{in}$  watts/m<sup>2</sup> impinging upon a flat shield as in Figure 8-2. When the wave encounters the first surface of the shield, a portion ( $P_{r1}$ ) of the

incident power is reflected back toward the source; the remainder ( $P_{t1}$ ) penetrates the shield and begins to propagate through the shield. The ratio of reflected power to incident power ( $P_{r1}/P_{in}$ ) depends upon the intrinsic impedance of the shield material and the wave impedance\* of the incident wave in the same manner as at the junction of two transmission lines of different characteristic impedances. A portion of the power transmitted into the shield ( $P_{t2}$ ) is converted into heat as the wave moves through the shield; this energy loss is referred to as absorption loss and is analogous to the dissipated energy within a lossy transmission line. Of the power which propagates through the shield to reach the second surface of the shield, a portion is reflected back into the shield and the remainder ( $P_{out}$ ) is transmitted through the surface and beyond the shield. If the absorption loss within the shield is small (less than 10 dB), a significant part of the power reflected at the second surface ( $P_{r2}$ ) propagates back to the first surface where a portion is reflected back into the shield, propagates back to-and-through the second surface, and contributes to the power propagated beyond the shield. Shielding effectiveness, then, depends upon three factors: (1) reflection loss, (2) absorption loss, and (3) a re-reflection factor which is significant when the absorption is small.

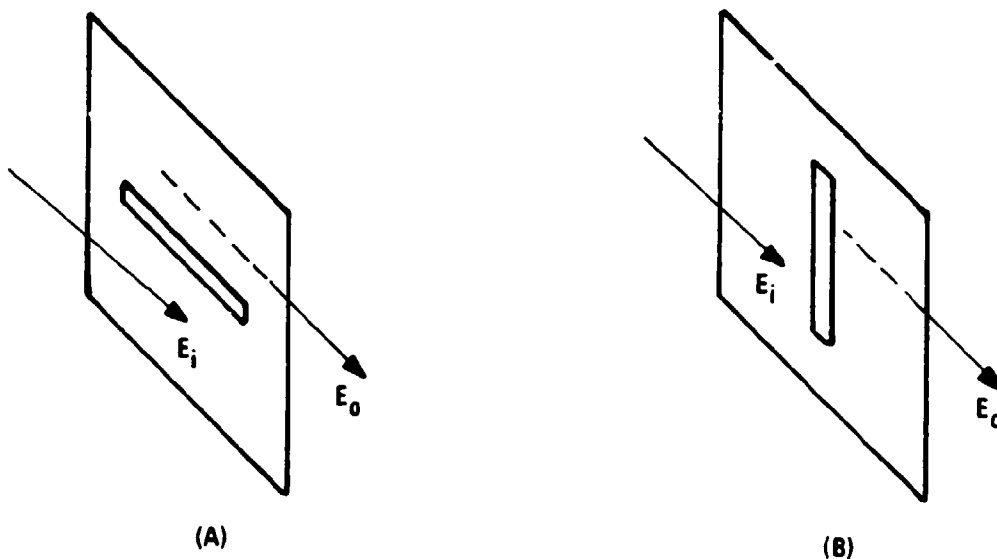


Figure 8-1. Electromagnetic Transmission Through a Slot

\*Wave impedance is defined as the ratio of the electric field strength to the magnetic field strength in the plane of interest. For further information, see Chapter 18 of Everett (8-2).

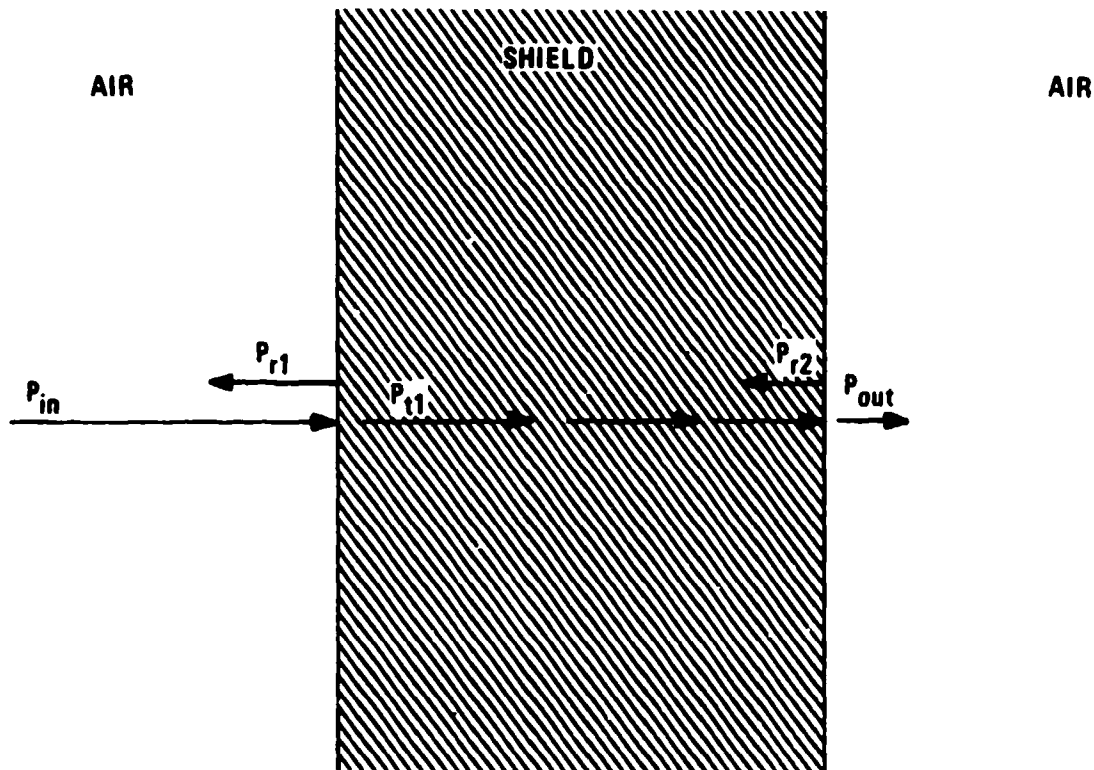


Figure 8-2. Transmission Line Model of Shielding

**8.2.3 Nonuniform Shielding.** Nonuniform shielding theory has been developed to deal with wave transmission through defects. It treats the defect as a transmission path in parallel with that representing transmission through the shielding material itself. The net shielding effectiveness of any practical enclosure is calculated as the result of all such parallel transmission paths, carefully considering transmission phase differences. The equipment design process, regardless of the theory utilized, consists of establishing undesired signal levels on one side of the proposed shielding barrier, estimating tolerable signal levels on the other side, and trading off shield design options to achieve the necessary effectiveness level.

### 8.3 SHIELDING EFFECTIVENESS OF CONTINUOUS SINGLE-THICKNESS SHIELDS.

The plane wave theory (or transmission line theory) of shielding is the basis of the most commonly used shielding design data. The resulting set of design equations, graphs, tables, and nomographs is based upon the separation of the shielding effectiveness into three additive terms: absorption loss, reflection loss, and a correction term to account for re-reflections within the shield.

The shield effectiveness (in decibels) of a large, plane sheet of metal with an EM wave arriving along a path perpendicular to the sheet has been shown (8-2) to be:

$$SE = \underset{A}{20 \log |e^{\gamma \ell}|} + \underset{R}{20 \log \left| \frac{1}{\tau} \right|} + \underset{C}{20 \log |1 - \Gamma e^{2\gamma \ell}|}, \quad (8-3)$$

where

$\ell$  = thickness of the shield,

$\gamma$  = propagation constant of the shield,

$\tau$  = transmission coefficient,

and

$\Gamma$  = reflection coefficient.

The shielding equation is often written as

$$SE = A + R + C \quad (8-4)$$

where A, R, and C are the indicated three terms in Equation 8-3 and represent respectively the Absorption Loss, the Reflection Loss, and the Correction Term for re-reflections as discussed earlier. In a particular shielding application, the values of the constants  $\gamma$ ,  $\Gamma$ , and  $\tau$  depend upon the conductivity ( $\sigma$ ), permeability ( $\mu$ ), and permittivity ( $\epsilon$ ) of the shielding material. The values of  $\Gamma$  and  $\tau$  depend also upon the wave impedance of the EM wave impinging upon the shield.

For convenience in the use of the shielding effectiveness equation, the individual terms A, R, and C have been expressed in more readily usable forms as functions of the EM wave's frequency ( $f$ ) and of the shield's thickness ( $\ell$ ), relative permeability ( $\mu_r$ ), and conductivity relative to copper ( $\sigma_r$ ). Simplified approximate expressions have been derived for the reflection and correction terms. The selection of the appropriate approximate expression will depend upon whether the wave impedance is low ( $Z_w < 377\Omega$ ; magnetic field), medium ( $Z_w \approx 377\Omega$ ; plane wave), or high ( $Z_w > 377\Omega$ ; electric field). Low impedance fields are found in the proximity of loop antennas, high impedance fields are found near dipole antennas, and plane waves exist away from the near fields of source antennas.

### 8.3.1 Absorption Loss.

The absorption loss of an EM wave passing through a shield of thickness  $\ell$  can be shown (8-2) to be given by:

$$A = K_1 \ell \sqrt{f \mu_r \sigma_r} \quad \text{dB}, \quad (8-5)$$

- where  $K_1 = 131.4$  if  $l$  is expressed in meters,  
 $3.34$  if  $l$  is expressed in inches,
- $l =$  shield thickness,
- $f =$  wave frequency, Hz,
- $\mu_r =$  relative permeability of shield material, and
- $g_r =$  conductivity of shield material relative to copper.

Note that the absorption loss (in decibels) is proportional to the thickness of the shield and also that it increases with the square root of the frequency of the EM wave to be shielded against. As to the selection of the shielding material, the absorption loss is seen to increase with the square root of the product of the relative permeability and conductivity (relative to copper) of the shield material.

Table 8-1 contains a tabulation of electrical properties of shielding materials ( $g_r$  and  $\mu_r$ ); since  $\mu_r$  is frequency dependent for magnetic materials, it is given for a typical shielding frequency of 150 kHz. The last two columns of Table 8-1 evaluate Equation 8-5 to give the absorption loss at 150 kHz for both a one millimeter and a one mil (0.001 inch) thick sheet for each of the listed materials. The absorption loss for other thickness can be calculated by simply multiplying by the shield thickness in millimeters or mils. Shield thicknesses are commonly expressed in either millimeters (mm) or mill-inches (mils); these two units are related as follows:

$$1 \text{ mm} = 39.37 \text{ mils or } 1 \text{ mil} = 0.0254 \text{ mm}$$

The variation of absorption loss with frequency, as well as a comparison of the absorption loss of three common shielding materials one mm thick, can be seen in Table 8-2. Also included is a listing of the relative permeability, as a function of frequency, for iron. Figure 8-3 presents the data of Table 8-2 in graphical form.

Remember that the absorption loss is just one of three additive terms which combine to give the attenuation (shielding efficiency) of the shield. At this point, the absorption loss has been presented in equation form (Equation 8-5), tabular form (Tables 8-1 and 8-2), and graphical form (Figure 8-3). The tabular and graphical forms are easy-to-use sources of accurate results when the shield material and frequency of interest are included in those tables and graphs. Quick results for almost any material and frequency combination can be obtained from an absorption nomograph (see Vol II), but the results are generally less precise; nomographs are a good source of data for initial design purposes. Once a shielding material and thickness are tentatively selected, one may wish to compute a more precise value of the absorption loss by evaluation of Equation 8-5.

### 8.3.2 Reflection Loss.

According to Equation 8-3, the reflection loss portion,  $R$ , of the shielding effectiveness,  $SE$ , is given by:

$$R = -20 \log |\tau| \text{ dB}, \quad (8-6)$$

where  $\tau$  is the transmission coefficient for the shield. The reflection loss includes the reflections at both surfaces of the shield (see Figure 8-2) and is dependent upon the wave impedance and frequency of the impinging EM wave as well as upon the electrical parameters of the shielding material. It is independent of the thickness of the shield.

Table 8-1

Electrical Properties of Shielding Materials at 150 kHz (8-3)

Metal	Relative Conductivity $\sigma/\sigma_c$	Relative Permeability $\mu/\mu_0$	Absorption Loss (dB)	
			1 mm thick	1 mil thick
Silver	1.05	1	51.96	1.32
Copper, annealed	1.00	1	50.91	1.29
Copper, hard-drawn	0.97	1	49.61	1.26
Gold	0.70	1	42.52	1.08
Aluminum	0.61	1	39.76	1.01
Magnesium	0.38	1	31.10	.79
Zinc	0.29	1	27.56	.70
Brass	0.26	1	25.98	.66
Cadmium	0.23	1	24.41	.62
Nickel	0.20	1	22.63	.58
Phosphor-bronze	0.18	1	21.65	.55
Iron	0.17	1,000	665.40	16.90
Tin	0.15	1	19.69	.50
Steel, SAE 1045	0.10	1,000	509.10	12.90
Beryllium	0.10	1	16.14	.41
Lead	0.08	1	14.17	.36
Hypernick	0.06	80,000	3484.00*	88.50*
Monel	0.04	1	10.24	.26
Mu-metal	0.03	80,000	2488.00*	63.20*
Permalloy	0.03	80,000	2488.00*	63.20*
Steel, stainless	0.02	1,000	224.40	5.70

\* With no saturation by incident field.



In a manner analogous to the classical equations (8-1) describing reflections in transmission lines, the shield reflection loss can be expressed as:

$$R = 20 \log \frac{1 + S/2}{4S} \text{ dB}, \quad (8-7)$$

where S is defined as the ratio of the wave impedance to the shield's intrinsic impedance and is analogous to the voltage standing wave ratio in transmission line practice. While the shield's intrinsic impedance is easily determined from the electrical properties of the shield material, the wave impedance is highly dependent upon the type and location of the EM wave source, as indicated in Figure 8-4.

In order to present practical methods for determination of the reflection loss, three separate classes of EM waves are considered and approximations for the reflection loss relationships applicable to the three classes are presented. Since wave impedance is the ratio of electric to magnetic field strengths, a predominantly magnetic field will have a low impedance and a predominantly electric field will have a high impedance. The three wave impedance classes to be considered are low, medium, and high and are commonly referred to as the magnetic, plane wave, and electric field, respectively.

Table 8-2

Absorption Loss, A, of 1 mm Metal Sheet (8-2)

Frequency	Iron		Copper		Aluminum	
	$\mu_r$	A	$\mu_r$	A	$\mu_r$	A
		(dB)		(dB)		(dB)
60.0 Hz	1,000	13	1	1	1	0.8
1.0 kHz	1,000	54	1	4	1	3.0
10.0 kHz	1,000	171	1	13	1	10.0
150.0 kHz	1,000	863	1	56	1	40.0
1.0 MHz	700	1,430	1	131	1	103.0
3.0 MHz	600	2,300	1	228	1	178.0
10.0 MHz	500	3,830	1	416	1	325.0
15.0 MHz	400	4,200	1	509	1	397.0
100.0 MHz	100	5,420	1	1,310	1	1,030.0
1.0 GHz	50	12,110	1	4,160	1	3,250.0
1.5 GHz	10	6,640		5,090	1	3,970.0
10.0 GHz	1	5,420	1	13,140	1	10,300.0

Relative Conductivity,  $g_r$ : Iron - 0.17, Copper - 1.0, Aluminum - 0.61.

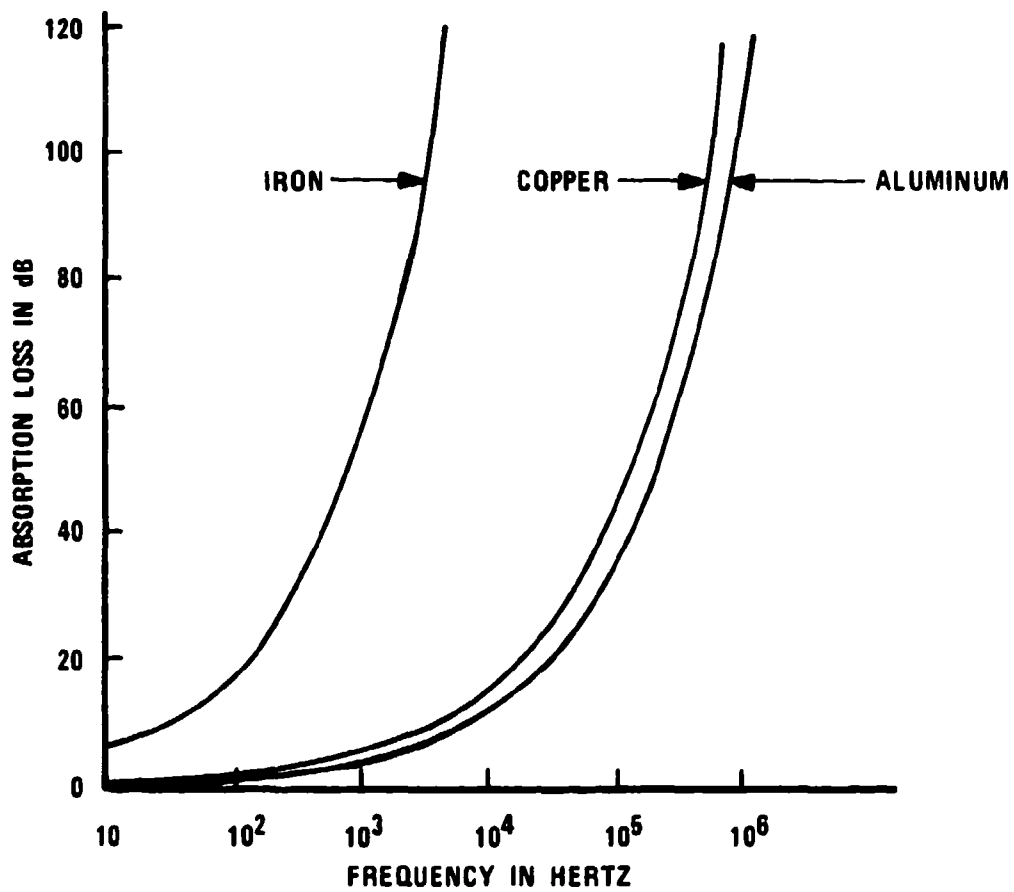
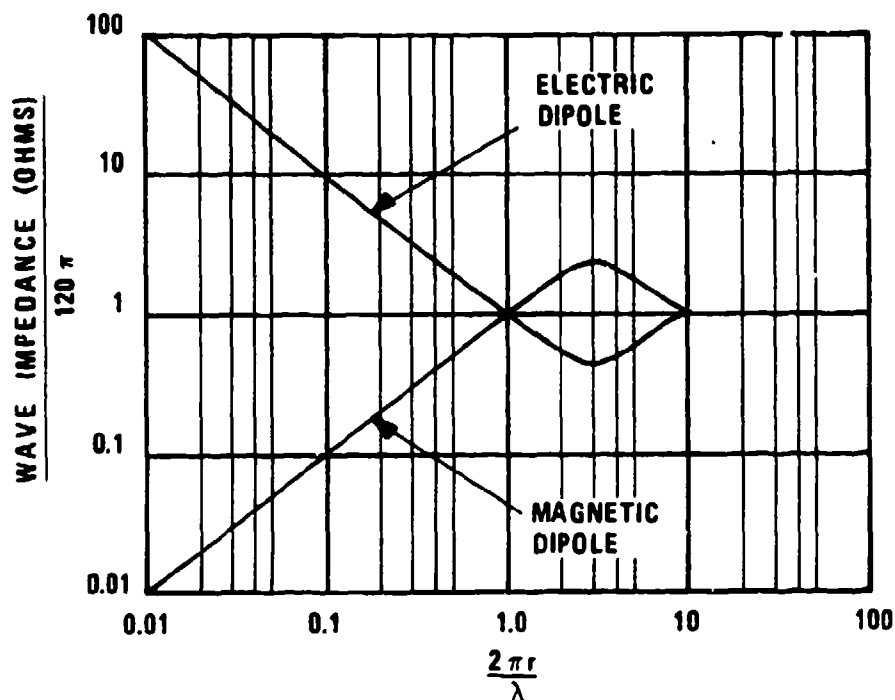


Figure 8-3. Absorption Loss for One Millimeter Shields



$r$  = DISTANCE BETWEEN SOURCE AND MONITORING POINT, IN SAME UNITS AS WAVELENGTH,  $\lambda$

Figure 8-4. Wave Impedance versus Distance from Source

#### 8.3.2.1 Low Impedance Field.

A loop, or magnetic dipole, antenna produces an EM wave which is predominantly magnetic in the near field ( $r \leq \lambda/2\pi$ ), where  $r$  is the distance from the antenna and  $\lambda$  is the wavelength of the EM field. For such magnetic (low impedance) EM fields, the reflection loss can be approximated as follows:

$$R_M = 20 \log \left( \frac{C_1}{r\sqrt{f g_r / \mu_r}} + C_2 r\sqrt{f g_r / \mu_r} + 0.354 \right), \quad (8-8)$$

where

- $r$  = distance from EM source to shield,  
 $f$  = frequency (Hz),  
 $g_r$  = conductivity of shield material relative to copper,  
 $\mu_r$  = relative permeability of shield material,

and the constants  $C_1$  and  $C_2$  depend upon the choice of units for the distance,  $r$ , as given in Table 8-3.

Table 8-3

## Coefficients for Magnetic Field Reflection Loss

<u>Coefficient</u>	<u>Units for Distance (r)</u>			
	<u>Meters</u>	<u>Millimeters</u>	<u>Inches</u>	<u>Mils</u>
$C_1$	.0117	11.7	0.462	462
$C_2$	5.35	.0053	0.136	136

As with absorption loss, the reflection loss for low impedance fields depends upon electrical properties of the shield material and upon the EM wave frequency. However, the reflection loss depends upon the distance from the source to the shield rather than upon the shield thickness.

Figure 8-5 shows the reflection loss as a function of frequency for iron, copper, and aluminum shields at distances of one inch (2.54 cm) and ten inches (25.4 cm) from the low impedance EM field source. For a given separation distance, the reflection loss is seen to be greater for copper and aluminum than for iron except at the lower frequencies where iron has a better reflection loss. The curves cannot be extended to higher frequencies for these separation distances since the approximations used in the derivation of Equation 8-8 assume that the separation distance,  $r$ , is less than  $\lambda/2\pi$ . For higher frequencies at these distances, the EM fields are more closely approximated by plane waves rather than by low impedance fields. Figure 8-6 is a universal curve of the reflection loss for low-impedance sources as a function of the parameter

$$0.394 \sqrt{\frac{fg_r}{\mu_r}} r.$$

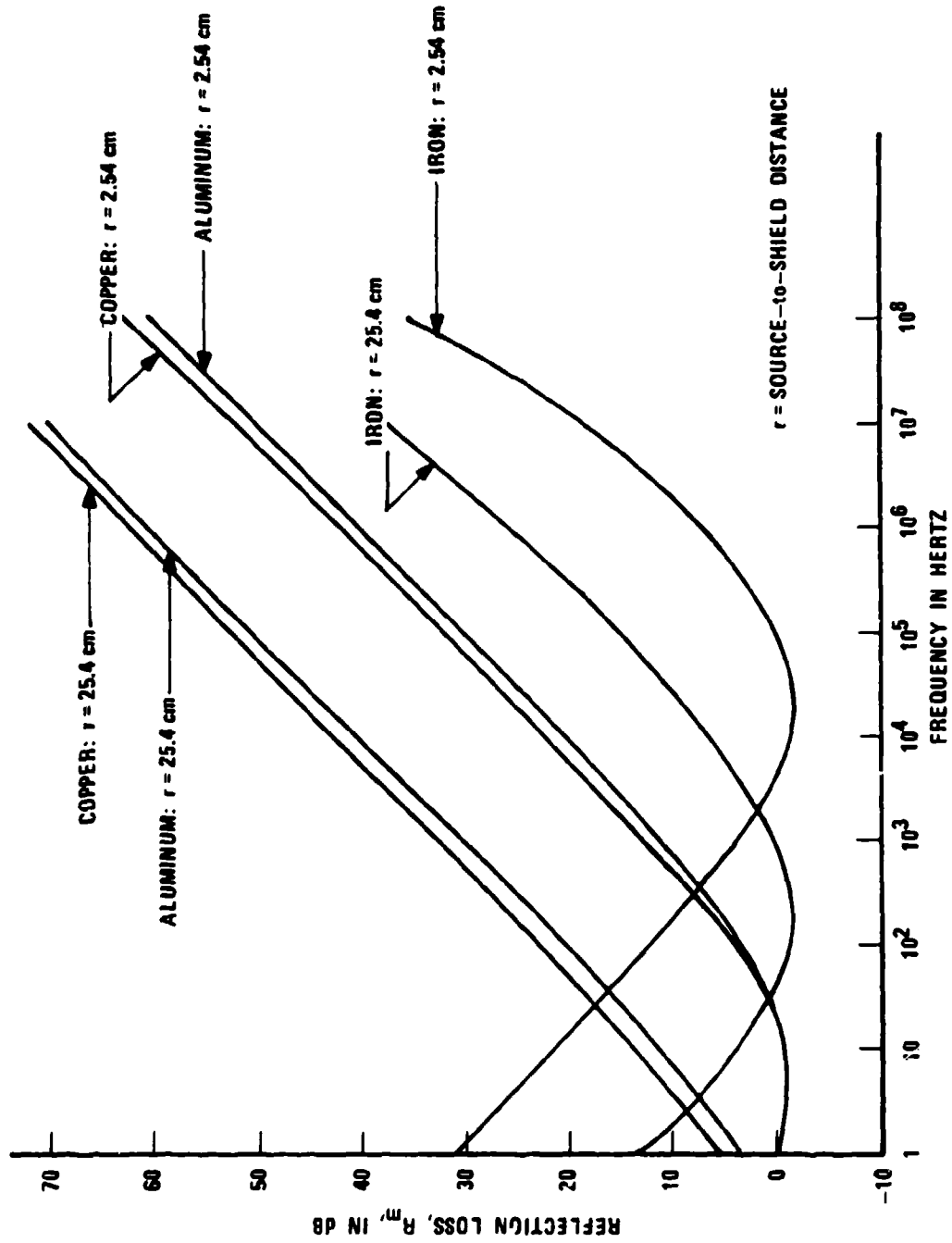


Figure 8-5. Reflection Loss for Iron, Copper, and Aluminum with a Low Impedance Source

### 8.3.2.2 Plane Wave Field.

The EM field at a distance of more than a few wavelengths from its source is essentially a plane wave with a wave impedance equal to the intrinsic impedance of the propagation media ( $377 \Omega$  for air). A plane wave has its electric field and magnetic field vectors,  $\underline{E}$  and  $\underline{H}$ , perpendicular both to each other and to the direction of propagation.

Unlike the low- and high-impedance fields associated with the near-fields of magnetic dipole and electric dipole sources, the plane wave field reflection loss is independent of the distance between the source and shield. The reflection loss for a plane wave impinging upon a uniform shield is given by

$$R_p = 168 - 20 \log \sqrt{\frac{f \mu_r}{g_r}} \quad (8-9)$$

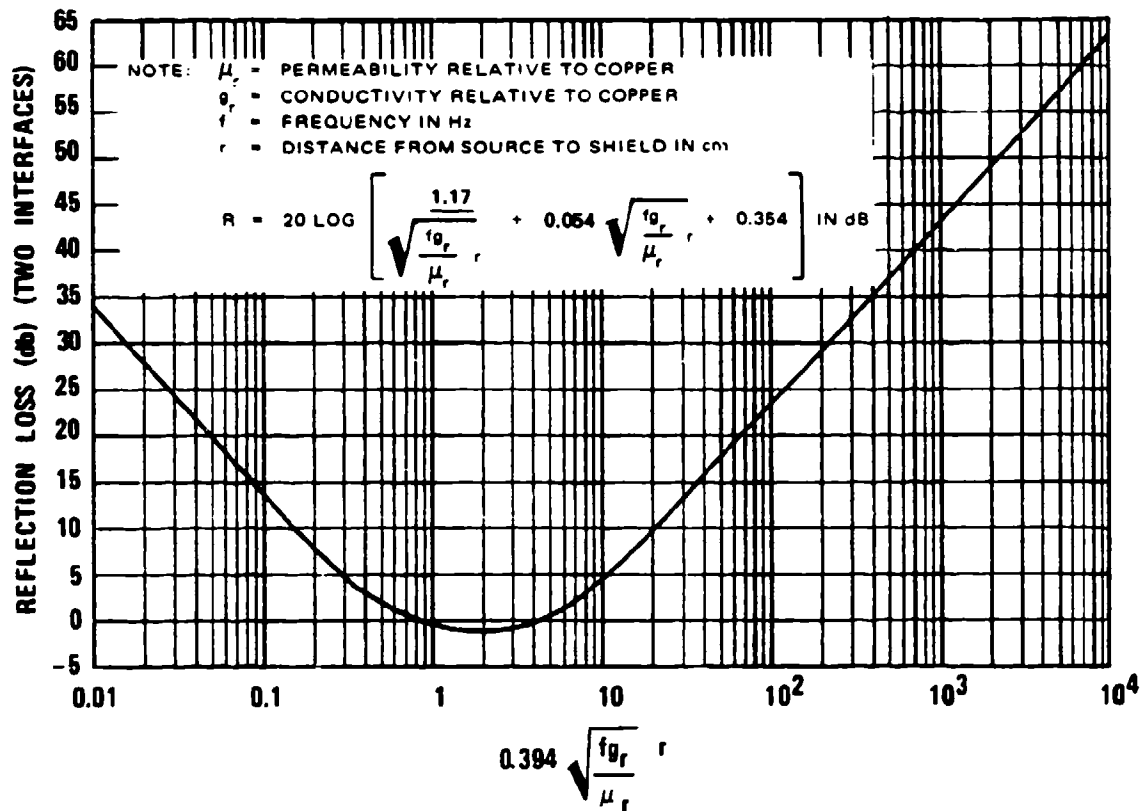
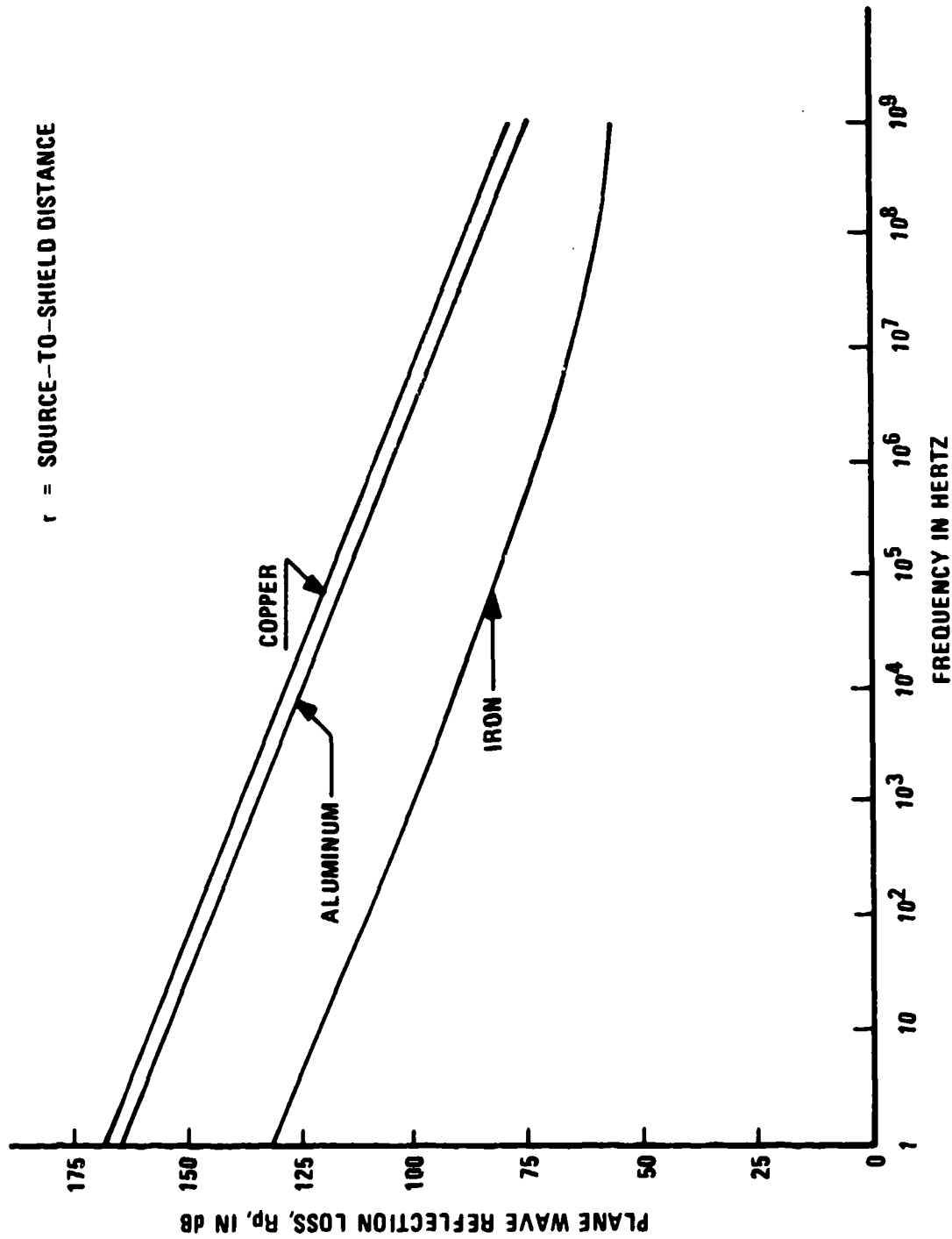


Figure 8-6. Universal Reflection Loss Curve for a Low Impedance Source (8-3)

Figure 8-7. Plane Wave Reflection Loss for Iron, Copper, and Aluminum ( $r > 2\lambda$ )

where  $g_r$ ,  $\mu_r$ , and  $f$  are as defined with Equation 8-8. The plane wave reflection loss is seen to decrease as the wave frequency increases, and to be better for shielding materials with lower  $\mu_r/g_r$  ratios. Figure 8-7 shows the plane wave reflection loss as a function of frequency for iron, copper, and aluminum shields. The curve for iron, unlike those for copper and aluminum, is not a straight line because iron's relative permeability is frequency dependent. Figure 8-8 provides a universal curve for plane wave reflection loss as a function of the parameter  $\sqrt{f\mu_r/g_r}$ .

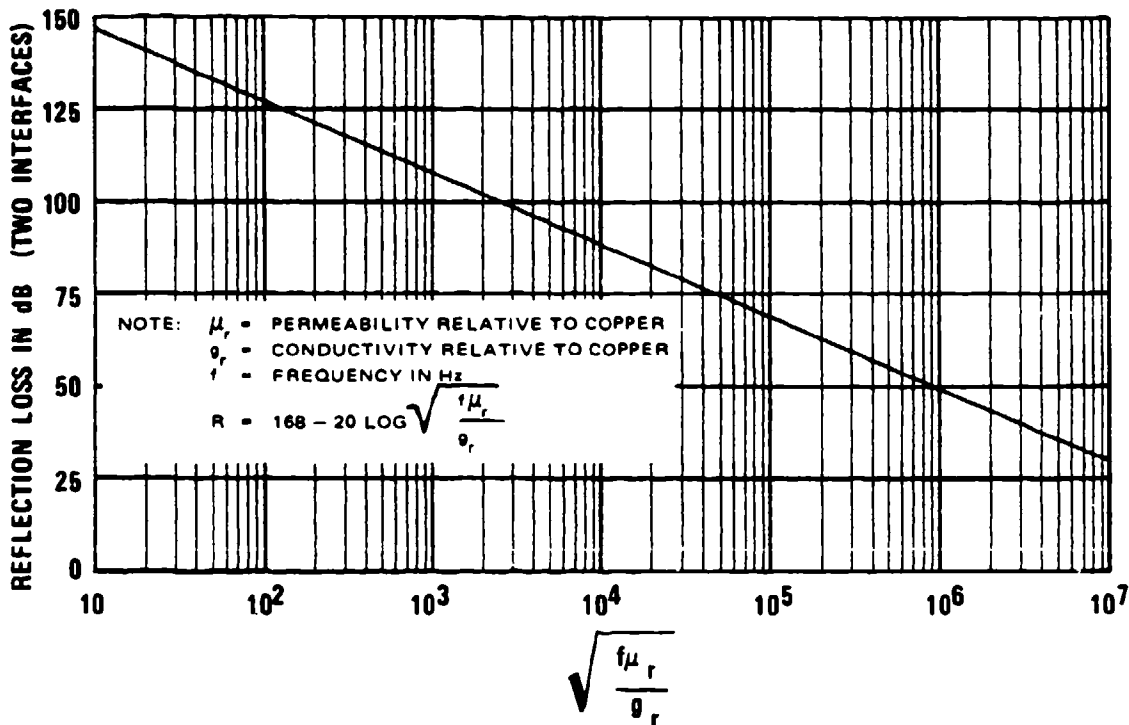


Figure 8-8. Universal Reflection Loss Curve for Plane Waves (8-3)

### 8.3.2.3 High Impedance Field.

The EM field in the proximity of an electric dipole antenna has a high electric field-to-magnetic field strength ratio (high wave impedance). The reflection loss for such a field encountering a shield is given by

$$R_E = C_3 - 10 \log \frac{\mu_r f^3 r^2}{g_r}, \quad (8-10)$$

where

$$\begin{aligned} C_3 &= 322, \text{ if } r \text{ is in meters} \\ C_3 &= 354, \text{ if } r \text{ is in inches,} \end{aligned}$$



and  $r$ ,  $g_r$ ,  $\mu_r$ , and  $f$  are as identified as in Equation 8-8. The high impedance EM wave reflection loss is seen to depend upon the separation distance,  $r$ , between the EM source and the shield, as does the low impedance case. The reflection loss is seen to decrease as the frequency increases and to be better when the ratio  $g_r/\mu_r$  is higher. Figure 8-9 is a universal curve for the high impedance reflection loss; the upper line is for the parameter range.

$$1 < 0.394r\sqrt{\frac{\mu_r f^3}{g_r}} < 10^6$$

and the lower line covers the range

$$0.394r\sqrt{\frac{\mu_r f^3}{g_r}} > 10^6$$

Figure 8-10 shows a plot of the high impedance EM wave reflection loss as a function of frequency for iron, copper, and aluminum for source-to-shield separation distances of one and ten inches. Separate curves for copper and aluminum are not shown since the high impedance reflection loss for aluminum is only 2 dB below that of copper.

The reflection losses for iron, copper, and aluminum shields at representative frequencies for magnetic, electric, and plane waves are given in Table 8-4. The source-to-shield distance for the magnetic and electric wave cases is one foot (30.5 cm).

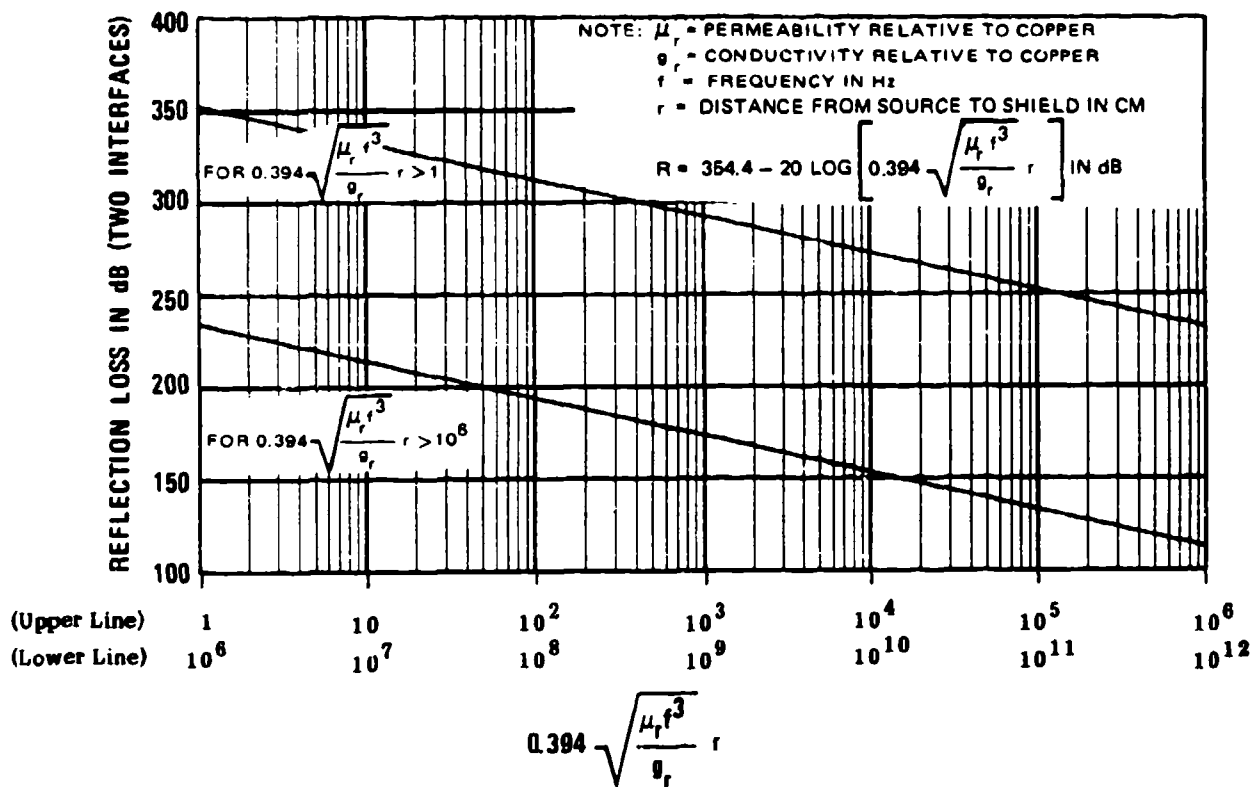


Figure 8-9. Universal Reflection Loss Curve for High Impedance Fields (8-3)

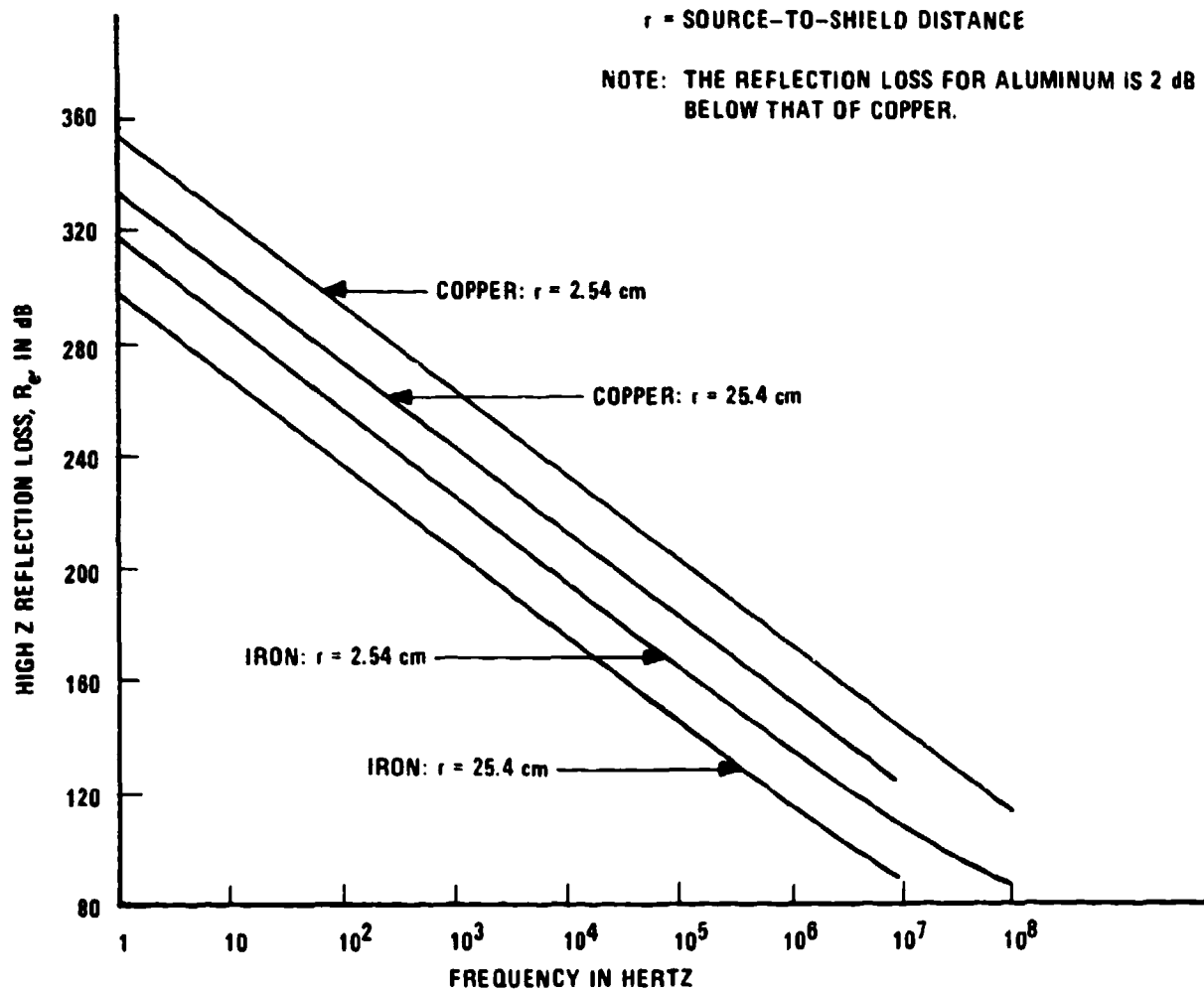


Figure 8-10. Reflection Losses for Iron, Copper, and Aluminum with a High Impedance Source

Table 8-4  
Calculated Reflection Loss in dB of Metal Sheet, Both Faces (8-2), (8-3)

Frequency	Reflection Loss*							
	Iron ( $g_r = 0.17$ )				Copper ( $\mu_r = 1, g_r = 1$ )			
	Magnetic Field $r = 30.5 \text{ cm}$		Electric Field $r = 30.5 \text{ cm}$		Plane Wave Field		Electric Field $r = 30.5 \text{ cm}$	
$\mu_r$	$r = 30.5 \text{ cm}$		$r = 30.5 \text{ cm}$				$r = 30.5 \text{ cm}$	
60.0 Hz	1000	-0.9 **	241	113	22	279	-	-
1.0 kHz	1000	0.9	204	101	34	242	-	-
10.0 kHz	1000	8.0	174	91	44	212	-	-
150.0 kHz	1000	10.0	139	79	56	177	54	175
1.0 MHz	700	28.0	116	72	64	152	62	150
15.0 MHz	400	42.0	83	63	76	117	74	115
100.0 MHz	100	57.0	64	61	84	92	82	90
1.5 GHz	10	59.0 +	59 +	59	76+	76 +	-	-
10.0 GHz	1	61.0 +	61 +	61	69+	68 +	-	-

\* If Absorption Loss < 15 dB, use Correction Factor C

\*\* The Reflection Loss for iron is zero at 620 Hz, it is a negative quantity at 60 Hz, zero at 31.5 Hz, and becomes positive at lower frequencies.

\*\*\* Plane-Wave Field not provided (normally close to that of copper).

+ At these frequencies and for given distance, the field becomes a plane wave.

### 8.3.3 Re-Reflection Correction Factor.

For shields in which the absorption loss (A) is reasonably large, say at least 10 dB, the energy reflected back into the shield at the second surface does not contribute significantly to the wave propagated through and beyond the shield. However, when the shield's absorption loss is low, a significant amount of energy is reflected at the second surface and finally propagated into the area to be shielded. Accordingly, for shields with low absorption, the shielding effectiveness is calculated as the sum of (1) the absorption loss, A, (2) the reflection loss, R, and (3) a re-reflection correction factor, C. The correction factor is

$$C = 20 \log [1 - \Gamma 10^{-A/10} (\cos [0.23A] - j \sin [0.23A])], \quad (8-11)$$

where A is the shield's absorption loss (see Equation 8-5) and  $\Gamma$  is the two-boundary reflection coefficient;  $\Gamma$  is dependent upon both the shield characteristic impedance and the wave impedance of the impinging EM wave. Equations for the reflection coefficient,  $\Gamma$ , are given in terms of a precalculation parameter, m, for each of three wave impedance classes in Table 8-5.

Values of the re-reflection correction terms for iron and copper sheets of various thicknesses and typical frequencies are given in Table 8-6. The correction term is seen to approach zero for thick shields or high frequencies since these conditions correspond to large absorption losses in the shield. The larger absorption loss of iron (compared with copper) for fixed frequency and thickness is also seen to result in a smaller correction term. Figure 8-11 presents the correction term in graphical form for copper in a magnetic (low impedance) field. Figure 8-12 presents a universal absorption loss curve (Equation 8-5). Recall that the correction term (Equation 8-11) depends upon the absorption loss, A, and that the reflection coefficient,  $\Gamma$ , is essentially unity. Whenever the approximation  $\Gamma = 1$  is valid, the correction term depends only upon the value of the absorption loss. For such conditions, the sum of the absorption loss and the re-reflection correction term is given by the dashed line on the universal curve in Figure 8-12.

**8.3.4 Total Shielding Effectiveness.** The item of interest for any shield is the (total) shielding effectiveness, i.e., the sum of the absorption loss (A), reflection loss (R), and the multi-reflection correction term (C). The terms, A, R, and C are of significance only as a means of predicting the shielding effectiveness. Table 8-7 contains the individual terms and the total shielding effectiveness for various shield thicknesses and EM wave frequencies for copper, iron, and aluminum shields. The entries under "SOURCE" designate the EM wave impedance classifications: L indicates a loop antenna and designates a predominantly magnetic field, D indicates an electric dipole antenna and designates a predominantly electric field, and P indicates a plane wave ( $Z_w = 377 \Omega$ ). All entries except the plane waves are for a source-to-shield separation distance of one foot. Figures 8-13 and 8-14 illustrate the total theoretical shielding performance which one may expect to obtain from enclosures constructed from copper foil and iron sheet to the electric, magnetic and plane wave propagation modes, although, the effect of doors, ventilation apertures, and power line penetrations has not been considered; in many applications these penetrations, together with techniques used for joining the shield materials, markedly reduce the overall practical insertion loss of a shielded enclosure.

Table 8-5  
Coefficients for Evaluation of Re-Reflection Correction Term, C

Type Field	Reflection Coefficient, $\Gamma$	Parameter, m (for r in meters)
Magnetic (Low Impedance)	$\frac{1}{4} \frac{(1-m)^2 - 2m^2 + j2\sqrt{2} m(1-m)^2}{[(1 + \sqrt{2} m)^2 + 1]^2}$	$\frac{4.7 \times 10^{-2}}{r} \sqrt{\frac{\mu_r}{f g_r}}$
Plane Wave* ( $F_w = 377\Omega$ )	$\frac{1}{4} \frac{(1-m)^2 - 2m^2 - j2\sqrt{2} m(1-m)^2}{[(1 + \sqrt{2} m)^2 + 1]^2}$	$9.77 \times 10^{-10} \sqrt{\frac{f \mu_r}{g_r}}$
	$= 1$	
Electric (High Impedance)	$\frac{1}{4} \frac{(1-m)^2 - 2m^2 - j2\sqrt{2} m(1-m)^2}{[(1 - \sqrt{2} m)^2 + 1]^2}$	$0.205 \times 10^{-16} r \sqrt{\frac{\mu_r f^3}{g_r}}$

\*For plane waves, m is so small that  $\Gamma$  is essentially unity.

Table 8-6

Correction Term C in dB for Single Metal Sheet (8-2)

Thickness* (mils)	Frequency					
	60 Hz	100 Hz	1 kHz	10 kHz	100 kHz	1 MHz
Copper, $\mu_r = 1$ , $g_r = 1$ , near field of loop						
1	-22.2	-24.00	-28.00	-20.0	-10.0	-3.0
5	-21.7	-22.00	-18.00	-7.0	-0.6	+0.1
10	-19.2	-19.00	-10.00	-3.0	+0.6	0.0
20	-15.6	-14.00	-5.00	+0.1	+0.1	
30	-13.0	-11.00	-3.00	+0.6	0.0	
50	-9.0	-7.00	-0.60	0.0		
100	-4.0	-3.00	+0.50			
200	-0.8	+0.50	0.00			
300	+0.3	+0.50				
Copper, $\mu_r = 1$ , $g_r = 1$ , plane waves and near field of electric dipole						
1	-42.0	-40.00	-30.00	-20.0	-10.0	-3.0
5	-28.0	-25.00	-18.00	-7.0	-0.6	+0.1
10	-22.0	-20.00	-10.00	-3.0	+0.6	0.0
20	-16.0	-14.00	-5.00	+0.1	+0.1	
30	-13.0	-11.00	-3.00	+0.6	0.0	
50	-9.0	-7.00	-0.60	+0.1		
100	-4.0	-3.00	+0.50	0.0		
200	-0.6	+0.10	0.00			
300	+0.1	+0.60				
Iron, $\mu_r = 1000$ , $g_r = 0.17$ , near field of loop						
1	1.0	1.00	-1.60	-1.3		
5	0.9	0.90	-0.60	0.0		
10	0.8	0.50	+0.08			
20	0.4	0.08	0.00			
30	0.06	0.06				
50	0.00	0.00				
Iron, $\mu_r = 1000$ , $g_r = 0.17$ , plane waves and near field of electric dipole						
1	-20.00	-17.00	-8.0	-1.3		
5	-7.00	-5.00	+0.2	0.0		
10	-3.00	-1.30	+0.4			
20	+0.20	+0.50	0.0			
30	+0.60	+0.40				
50	+0.10	0.00				

\*1 mil equals 0.0254 mm

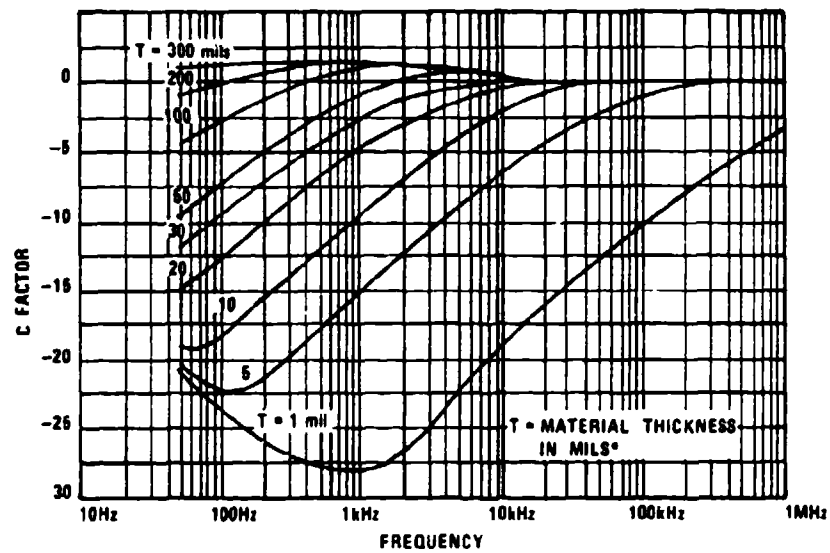


Figure 8-11. Graph of Correction Term (C) for Copper in a Magnetic Field (8-4)

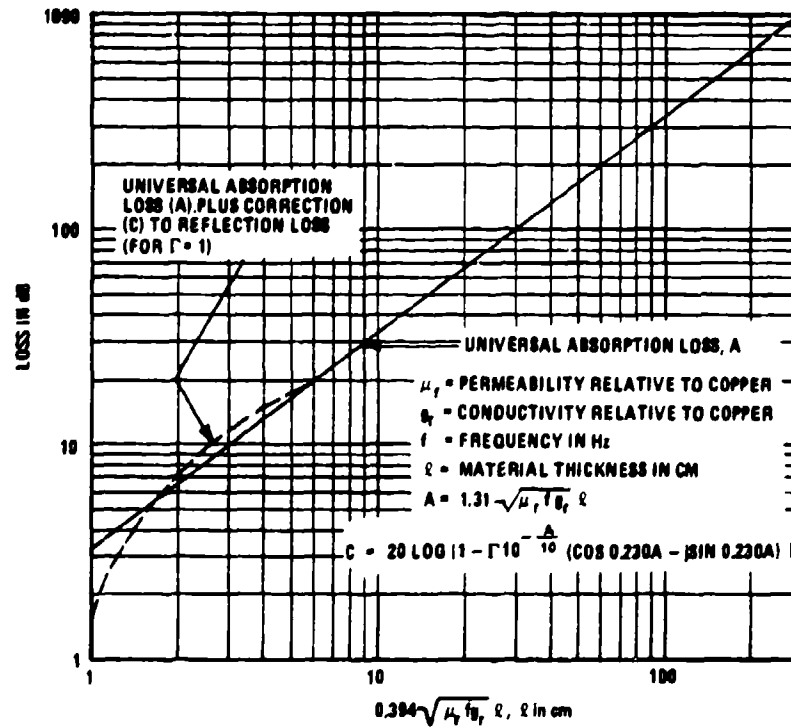


Figure 8-12. Absorption Loss and Multiple Reflection Correction Term when  $\Gamma = 1$  (8-2)

Table 8-7

## Calculated Values of Shielding Effectiveness (8-2)

<u>Thickness</u> (mils)	<u>Frequency</u> (Hz)	<u>Source*</u>	<u>R</u> (dB) <u>Copper</u>	<u>A</u> (dB)	<u>C</u> (dB)	<u>SE = A + R + C</u> (dB)
1	60	L	22.4	0.026	-22.20	0.23
10	60	L	22.4	0.260	-19.20	3.46
300	60	L	22.0	7.800	+0.32	30.12
10	1 k	L	34.2	1.060	-10.37	24.89
10	10 k	L	44.2	3.340	+2.62	50.16
10	10 k	D	212.0	3.340	-2.61	212.73
10	10 k	P	128.0	3.340	-2.61	128.73
30	10 k	L	44.2	10.020	+0.58	54.80
10	150 k	L	56.0	12.900	0.50	69.40
10	150 k	D	176.8	12.900	0.50	190.20
10	150 k	P	117.0	12.900	0.50	130.40
10	1 M	L	64.2	33.400	0.00	97.60
10	1 M	D	152.0	33.400	0.00	185.40
10	1 M	P	108.0	33.400	0.00	141.40
10	15 M	L	76.0	129.00	0.00	205.00
10	15 M	D	116.0	129.00	0.00	245.00
10	15 M	P	96.0	129.00	0.00	225.00
10	100 M	L	84.0	334.00	0.00	418.00
10	100 M	D	92.0	334.00	0.00	426.00
10	100 M	P	88.0	334.00	0.00	422.00



Table 8-7 (Continued)

## Calculated Values of Shielding Effectiveness

<u>Thickness</u> (mils)	<u>Frequency</u> (Hz)	<u>Source*</u>	<u>R</u> (dB) <u>Iron</u>	<u>A</u> (dB)	<u>C</u> (dB)	<u>SE = A + R + C</u> (dB)
1	60	L	-0.9	0.33	+0.95	0.38
10	60	L	-0.9	3.30	+0.78	3.18
300	60	L	-0.9	100.00	0.00	99.10
10	1 k	L	0.9	13.70	+0.06	14.66
10	10 k	L	8.0	43.50	0.00	51.50
10	10 k	D	174.0	43.50	0.00	217.50
10	10 k	P	99.5	43.50	0.00	143.00
30	10 k	L	8.0	130.50	0.00	138.50
10	150 k	L	19.0	160.00	0.00	179.00
10	150 k	D	139.0	160.00	0.00	308.00
10	150 k	P	79.0	160.00	0.00	248.00
10	1 M	L	28.0	363.00	0.00	391.00
10	1 M	D	116.0	363.00	0.00	479.00
10	1 M	P	72.0	363.00	0.00	435.00
10	15 M	L	42.0	1060.00	0.00	1102.00
10	15 M	D	83.0	1060.00	0.00	1143.00
10	15 M	P	63.0	1060.00	0.00	1123.00
10	100 M	L	56.0	1370.00	0.00	1426.00
10	100 M	D	64.0	1370.00	0.00	1434.00
10	100 M	P	60.0	1370.00	0.00	1430.00

Table 8-7 (Continued)

## Calculated Values of Shielding Effectiveness

<u>Thickness</u> (mils)	<u>Frequency</u> (Hz)	<u>Source*</u>	<u>R</u> (dB) <u>Aluminum</u>	<u>A</u> (dB)	<u>C</u> (dB)	<u>SE = A + R + C</u> (dB)
10	1 M	L	62.0	26.0	0.00	88.0
10	1 M	D	150.0	26.0	0.00	176.0
10	1 M	P	-	-	-	-
10	15 M	L	79.0	100.0	0.00	179.0
10	15 M	D	115.0	100.0	0.00	215.0
10	15 M	P	-	-	-	-
10	100 M	L	82.0	260.0	0.00	342.0
10	100 M	D	90.0	260.0	0.00	350.0
10	100 M	P	-	-	-	-

\*L = near field of loop or magnetic dipole,  $r = 30$  cm from shield.

D = near field of electric dipole,  $r = 30$  cm from shield.

P = plane wave.

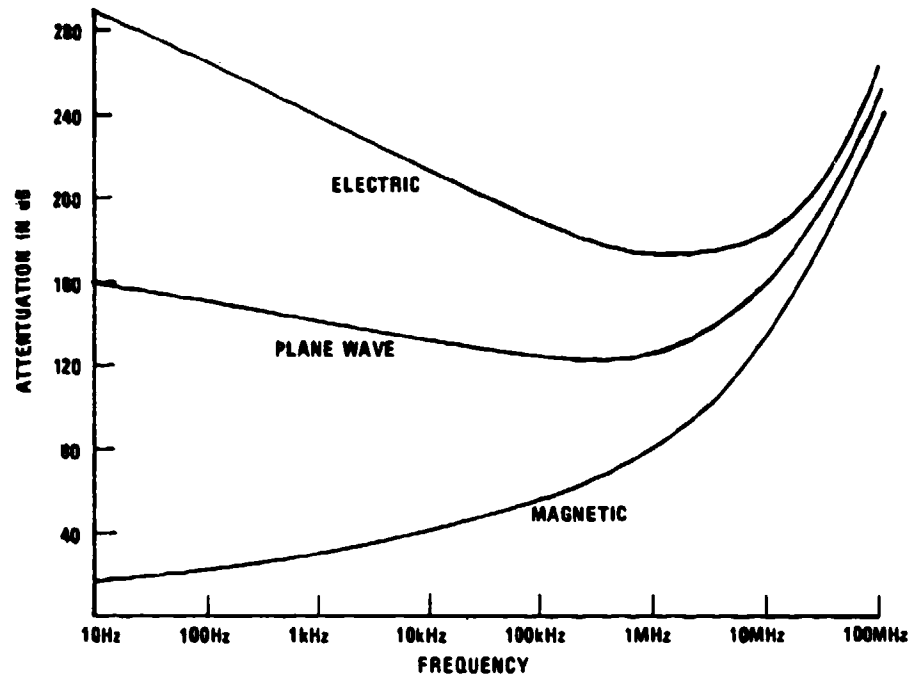


Figure 8-13. Theoretical Attenuation of Thin Copper Foil (8-5)

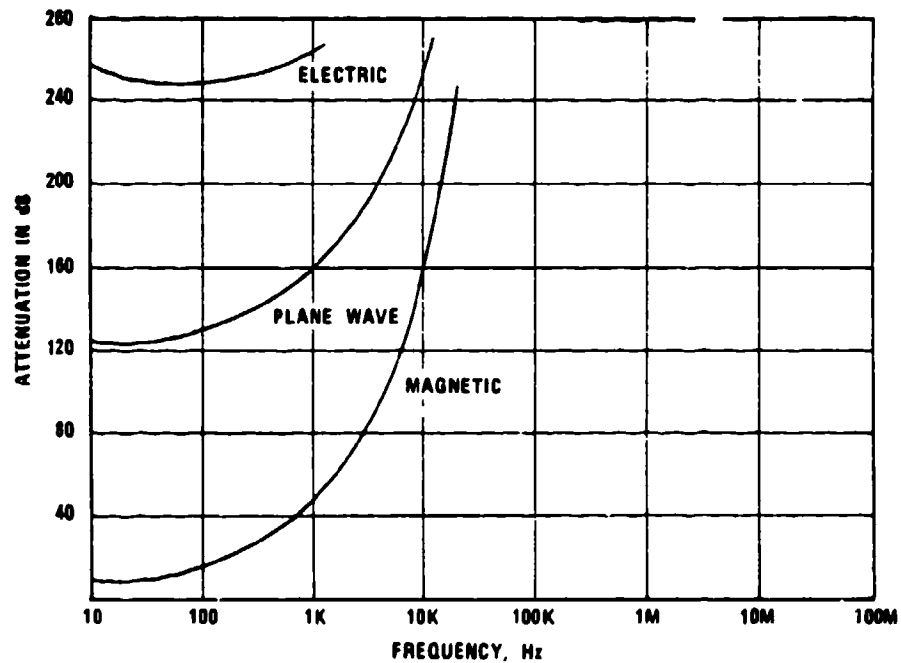


Figure 8-14. Theoretical Attenuation of Thin Iron Sheet (8-5)

#### 8.3.4.1 Measured Data.

In contrast to the theoretical shielding effectiveness presented thus far, Table 8-8 and Figures 8-15 and 8-16 present actual measured data. Figure 8-15 illustrates representative shielding effectiveness data taken for a variety of high-permeability sheet materials. Loop sensors were located 0.3 cm (1/8") from each sheet. The figure shows the typical leveling off in shielding effectiveness as frequency is decreased, with the breakpoint occurring in the 1-kHz range. Low frequency magnetic shielding is essentially achieved by establishing a low reluctance path in which the magnetic field is contained. The variation of shielding effectiveness as a function of loop sensor separation is shown in Figure 8-16 for one of the materials plotted in the previous figure. A change in effectiveness of about 5 dB over the range of the test at a particular frequency is indicated.

A difficulty with most magnetic shielding materials is their tendency to change permeability when formed, machined, subjected to rapid or extreme temperature changes, or dropped. These processes change the orientation of the magnetic domains in the material, and it is necessary to reorient the domains by annealing to restore the initial magnetic properties. A typical annealing process involves heating the material to about 2000° F (sometimes in an inert gas environment), holding it at that temperature for approximately two hours, and letting it slowly cool to room temperature.

#### 8.3.4.2 Summary.

The shielding effectiveness in dB for a shield is calculated as the sum of three terms: absorption loss (A), reflection loss (R), and a correction term (C). The absorption loss is independent of the distance from the EM source. It depends upon the shield thickness and the shielding material's conductivity and permeability, as well as upon the frequency of the incident EM wave. However, the reflection loss (like that of a junction of two types of transmission lines) depends upon the ratio of the EM wave impedance to the shield impedance and is therefore dependent upon both the EM source type and the distance between the source and shield. It is also dependent upon the EM source frequency and the shield material's conductivity and permeability but does not depend upon the thickness of the shield. The multi-reflection correction term is essentially zero for shields with absorption losses greater than 10 dB; for shields with less absorption loss the correction factor should be used. It is dependent upon the EM wave impedance classification and the absorption loss, as well as the frequency, conductivity, and permeability. Table 8-9 summarizes the shielding equations.

Equations, tables, and graphs, have been presented for evaluation of the components of the shielding effectiveness. The choice of which form to use will be influenced by the time available to the user and the accuracy to which the data is needed.

Table 8-8

Measured Shielding Effectiveness in dB for Solid-Sheet Materials (8-3)

Type Wave	Material	Thickness (mils)	Nominal Effectiveness							
			0.1 kHz	1 kHz	10 kHz	85 kHz	200 kHz	2 MHz	5 MHz	10 MHz
Plane	Cu	2.5					109	106	114	
Plane	Al	5.0					107	109	118	
Plane	Stainless Steel	18.0					97	95	99	
Plane	Steel ( $\mu_r = 250$ )	4.5					105	99	101	
Plane	AA-Conetic Foil ( $\mu_r = 10,000$ )	3.5					97	130		
Low Impedance	Cu	25.0	8	22	58					
		63.0				97				
		31.0	4	11	29	59		120		
		4.5				34		55		
Low Impedance	Al	125.0	5	18	50					
		63.0	1	16	35	78				
		31.0	1	10	24					
Low Impedance	Steel	63.0	25	40	80					
		31.0	4	28	59	94		92		120
Low Impedance	Cu-Clad Steel Clad 2 Sides	31.0				107				

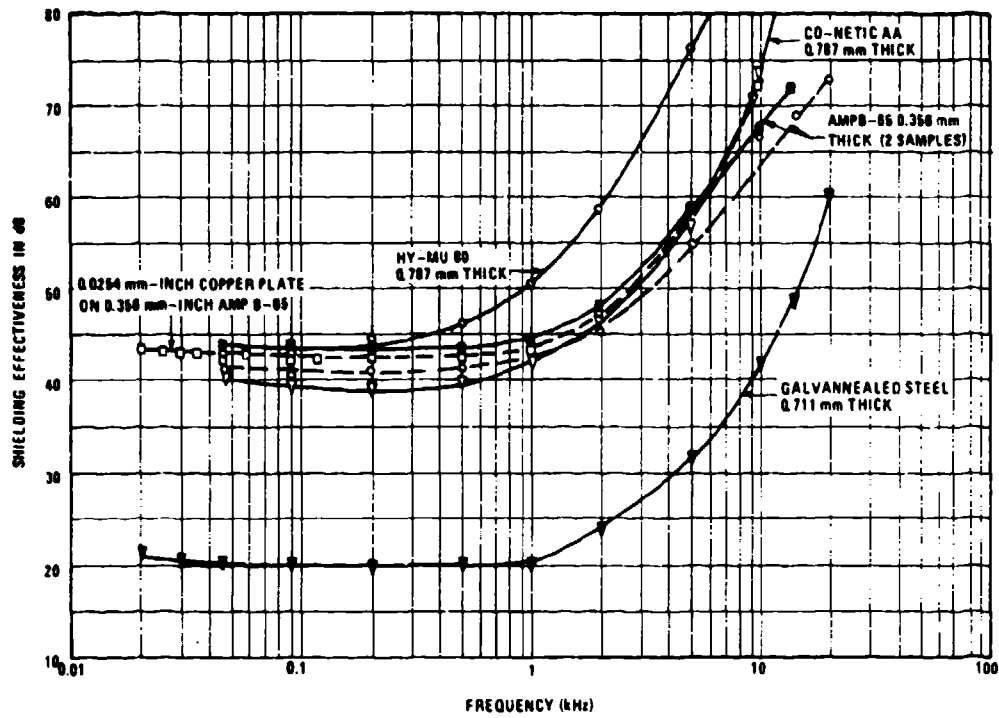


Figure 8-15. Measured Shielding Effectiveness of High Permeability Metals (8-6)

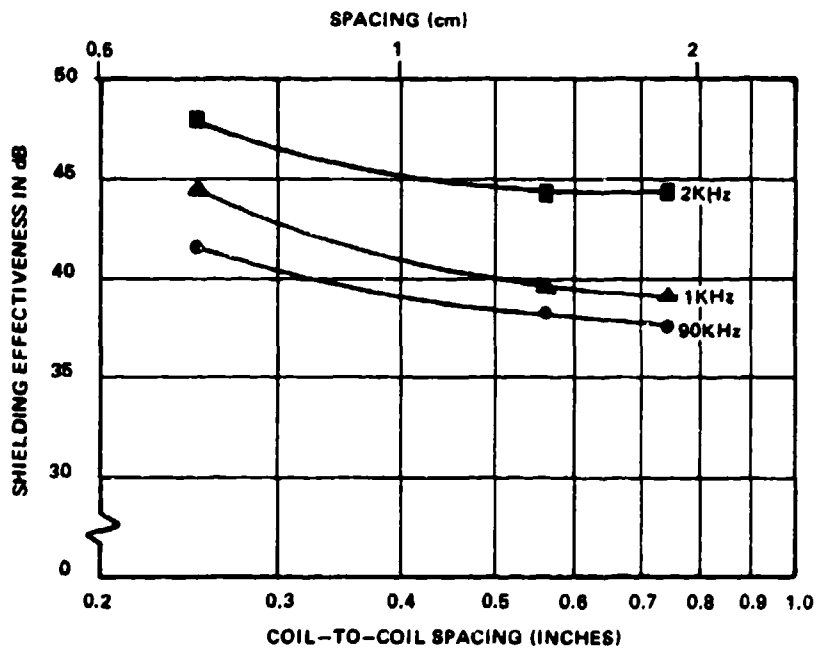


Figure 8-16. Measured Shielding Effectiveness of High Permeability Material as a Function of Measurement Loop Spacing (8-6)

Table 8-9

## Summary of Formulas for Shielding Effectiveness

$$SE = A + R + C$$

- $r$  = distance from source to shield (in meters)  
 $\mu_r$  = relative permeability  
 $g_r$  = conductivity relative to copper  
 $f$  = frequency (Hz)
- $k$  = antenna correction factor  
 $L$  = largest dimension of antenna (in meters)  
 $\Gamma$  = reflection coefficient  
 $m$  = precalculation parameter

Source	Attenuation, A	Reflection Loss, R	Correction Term C*
Plane Wave $r \geq \frac{kL^2}{\lambda}$ or $r \geq 2\lambda$ (dipole)	$131.41\sqrt{f\mu_r g_r}$	$168 + 20 \log \sqrt{\frac{g_r}{f\mu_r}}$	$\Gamma = 4 \frac{(1 - m^2)^2 - 2m^2 - 12\sqrt{2m}(1 - m^2)}{(1 + \sqrt{2m})^2 + 1} = 1$ $m = 9.77 \times 10^{-10} \sqrt{\frac{f\mu_r}{g_r}}$
Loop, near field	$131.41\sqrt{f\mu_r g_r}$	$-9 + 20 \log \frac{3.32 \times 10^{-2}}{r} \sqrt{\frac{\mu_r}{f g_r}} + 1 + 15.1r \sqrt{\frac{f g_r}{\mu_r}}$	$\Gamma = 4 \frac{(1 - m^2)^2 - 2m^2 + 12\sqrt{2m}(1 - m^2)}{(1 + \sqrt{2m})^2 + 1} = 2$ $m = \frac{4.7 \times 10^{-2}}{r} \sqrt{\frac{\mu_r}{f g_r}}$
Electric dipole, near field	$131.41\sqrt{f\mu_r g_r}$	$322 + \log \frac{1}{r} \sqrt{\frac{g_r}{f^3}}$	$\Gamma = 4 \frac{(1 - m^2)^2 - 2m^2 - m2\sqrt{2m}(1 - m^2)}{(1 + \sqrt{2m})^2 + 1} = 2$ $m = 0.205 \times 10^{-16} r \sqrt{\frac{\mu_r f^3}{g_r}}$

\*C =  $20 \log \{ 1 - \Gamma 10^{-A/10} (\cos 0.23A - j \sin 0.23A) \}$

#### 8.4 SHIELDING EFFECTIVENESS OF OTHER SHIELDS.

##### 8.4.1 Multiple Solid Shields.

There are cases when it is appropriate to consider using two or even three layers of shielding material rather than a single sheet to obtain particular total shielding characteristics. The most frequently encountered circumstances are when good protection against both electric and magnetic fields is desired, although other situations also occur.

Although Mumetal and similar types of high-permeability alloys provide good shielding for low-frequency weak magnetic fields, they tend to be less effective under the saturating effects of high-level fields. Where magnetic shielding in strong signal environments is necessary, it is often desirable to use a multiple shield, where the outer material has a lower permeability but a higher saturation level than the inner material. Such a structure might be constructed with materials having the characteristics given in Table 8-10.

Table 8-10

#### Magnetic Material Characteristics

<u>Property</u>	<u>Inner Material (Co-Netic AA)</u>	<u>Outer Material (Netic S 3-6)</u>
Initial Permeability	20,000.00	300.0
Permeability at 0.02 tesla	80,000.00	500.0
Saturation Inductance (tesla)	0.75	2.2

The material thickness necessary would be dictated by the unexpected levels of external fields and the desired suppression.

When much of the usefulness of shielding is due to reflection loss, two or more layers of metal separated by dielectric materials and yielding multiple reflections, will provide greater shielding than the same thickness of metal in a single sheet. The separation of the two layers of metal is necessary to provide for the additional discontinuous surfaces. A similar advantage has been noted with magnetic sheet materials (see Figure 8-17).

For the special case where two metallic sheets of the same material and thickness are separated by an air space, the penetration and reflection losses are each twice of those of a single sheet. However, the correction factors differ from double the value of a single sheet. One term in the correction factor is negative over much of the frequency range.



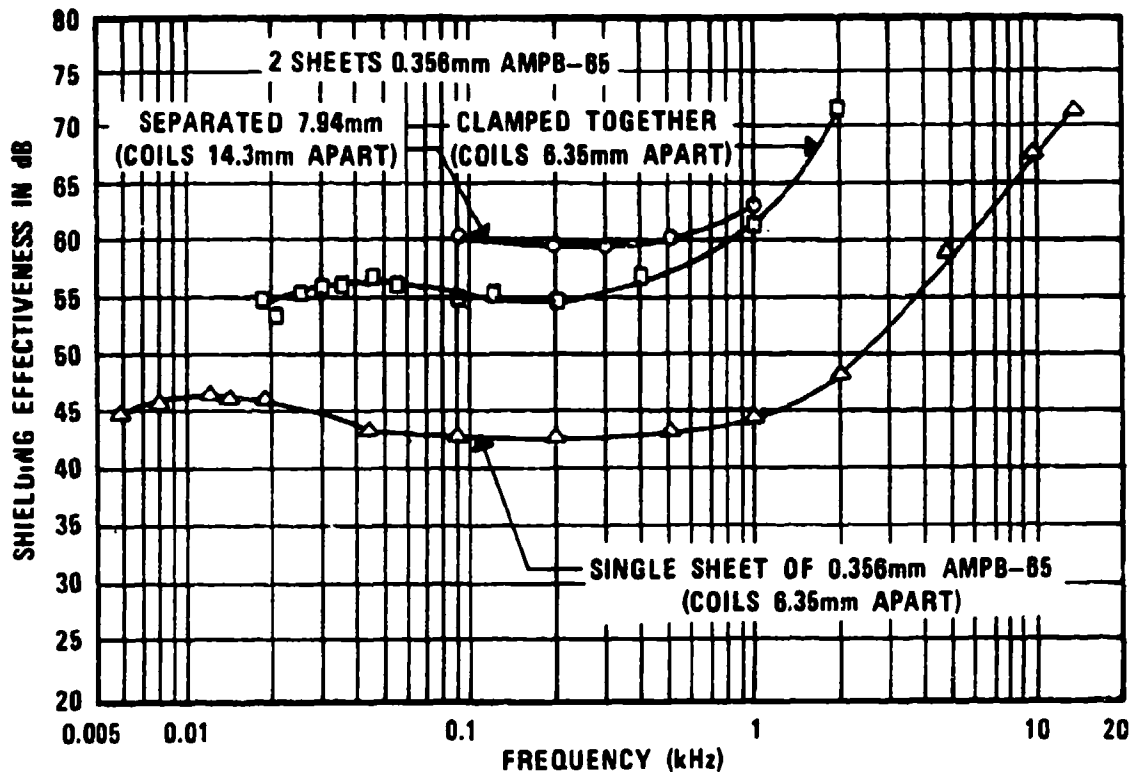


Figure 8-17. Measured Shielding Effectiveness of Two Sheets of a High Permeability Metal (8-6)

Consequently, a double shield is considerably less effective than the sum of two single shields. However, it is considerably more effective than a single shield of the same total thickness.

#### 8.4.2 Coatings and Thin-Film Shields.

Thin shielding\* has been employed in a variety of ways, ranging from metallized component packaging for protection against RF fields during shipping and storing, to vacuum deposited shields for microelectronics applications, and to wallpaper-like shielding material for shielded enclosures.

Solid material shielding theory is applicable to thin-film shields. For shields much thinner than  $\lambda/4$ , the absorption loss is very small, but the multiple reflection correction term  $C$ , is fairly large and negative, thus offsetting a portion of the reflection loss. The implication of the negative term is that the various reflections have additive phase relationships, and thus reduce the effectiveness of the shield. The shield effectiveness is essentially independent of frequency. When the shield thickness exceeds  $\lambda/4$ , the multiple reflection term becomes negligible, and there is no offsetting effect to the other losses. Thus the material shielding effectiveness increases and is frequency dependent.

\* The thickness of a thin-film shield is often expressed in Angstroms. This unit is related to mils by 1 Angstrom ( $\text{\AA}$ ) =  $3.937 \times 10^{-6}$  mils.

Table 8-11 provides representative calculations of the shielding effectiveness of thin-film cover for different thicknesses and frequencies. One-quarter wavelength in copper is approximately 0.13 mils at 1 GHz, and it can be seen that shield effectiveness changes significantly above this thickness.

Table 8-11  
Calculated Values of Copper Thin-Film Shielding Effectiveness in dB Against Plane-Wave Energy (8-7)

Thickness	0.0041 Mils		0.049 Mils		0.086 Mils		0.86 Mils	
Frequency	1 MHz	1 GHz	1 MHz	1 GHz	1 MHz	1 GHz	1 MHz	1 GHz
Absorption Loss, A	0.014	0.44	0.16	5.2	0.29	9.2	2.9	92
Single Reflection Loss, R	109	79	109	79	109	79	109	79
Multiple Reflection Correction Term, C	-47	-17	-26	-6	-21	0.6	-3.5	0
Shield Effectiveness, SE	62	62	83	78	88	90	108	171

#### 8.4.3 Screens and Perforated Metal Shields.

There are many applications in which the shield cannot be made of a solid material. Screens and perforated materials must be employed if an enclosure must be transparent (e.g., a meter face) or ventilated. The shielding effectiveness of solid metal shields has been treated from the viewpoint of classical transmission line theory in the preceding sections. To obtain an expression for shielding effectiveness which is applicable to screens and perforated metal sheets, it is necessary to account for the following:

- The attenuation effects of the individual shield apertures acting as many stacked waveguides-below-cutoff (see Section 8.5.3.1).
- Reflection losses, considering the geometry of the openings.
- Area of the opening when the test antenna is far from the shield in comparison to the distance between holes in the shield.
- Skin depth effects.
- Coupling between closely spaced openings.

The shielding effectiveness, in decibels, is expressed as follows (8-8)

$$SE = A_a + R_a + C_a + K_1 + K_2 + K_3 \quad (8-12)$$

where, as with solid shields,  $A_a$  represents the absorption or attenuation term,  $R_a$  the reflection loss term, and  $C_a$  the multi-reflection correction term. The additional terms  $K_1$ ,  $K_2$ , and  $K_3$  approximate the effects of items c, d, and e above. Detailed expressions for the screen and perforated metal sheet shielding effectiveness terms are given as follows for single layer wire cloth or screening:

$$\begin{aligned} A_a &= \text{aperture attenuation in dB,} \\ &= 27.3 D/W \text{ for rectangular apertures, and} \quad (8-13) \\ &= 32 D/d \text{ for circular apertures,} \quad (8-14) \end{aligned}$$

where

$$\begin{aligned} D &= \text{depth of aperture in inches,} \\ W &= \text{width of rectangular aperture in inches (measured perpendicular to the E-Vector),} \\ d &= \text{diameter of circular aperture in inches,} \\ R_a &= \text{aperture reflection loss in dB,} \\ &= 20 \log \left| \frac{(1+k)^2}{4k} \right|, \text{ and} \quad (8-15) \end{aligned}$$

$$\begin{aligned} C_a &= \text{correction factor for aperture reflections (negligible when } A_a \text{ is greater than 10 dB)} \\ &= 20 \log \left| 1 - \frac{(k-1)^2}{(k+1)^2} \times 10^{-A_a/10} \right| \quad (8-16) \end{aligned}$$

In Equations 8-15 and 8-16,

$$\begin{aligned} k &= \text{ratio of aperture characteristic impedance to incident wave impedance, or} \\ &= W/3.142r \text{ for rectangular apertures and magnetic fields} \quad (8-17) \end{aligned}$$

$$= d/3.682r \text{ for circular apertures and magnetic fields} \quad (8-18)$$

$$= jfW \times 1.7 \times 10^{-4} \text{ for rectangular apertures and radiated fields} \quad (8-19)$$

$$= jfd \times 1.47 \times 10^{-4} \text{ for circular apertures and radiated fields} \quad (8-20)$$

$$f = \text{frequency in MHz}$$

$$r = \text{distance from signal source to shield in inches}$$

$$j = \sqrt{-1}$$

$K_1$  = correction factor for number of openings per unit square (applicable when test antennas are far from the shield in comparison to distance between holes in the shield),

$$= 10 \log \frac{1}{an} \quad (8-21)$$

where

$a$  = area of each hole in square inches

$n$  = number of holes per square inch

$K_2$  = correction factor for penetration of the conductor at low frequencies

$$= -20 \log \left( 1 + \frac{35}{p^{2.3}} \right) \quad (8-22)$$

where

$p$  = ratio of the wire diameter to skin depth,  $\delta$ , where

$$\begin{aligned} \delta &= \frac{6.61}{\sqrt{f}} \text{ in cm, or} \\ &= \frac{2.60}{\sqrt{f}} \text{ in inches, } f \text{ in Hz} \end{aligned} \quad (8-23)$$

$K_3$  = correction factor for coupling between closely spaced shallow holes

$$= 20 \log \frac{1}{\tanh(A_a/8.686)} \quad (8-24)$$

As an example, determine the shielding effectiveness of a No. 22, 15 mil copper screen when subjected to a predominantly magnetic field from a loop source 1.75 inches away and operating at a frequency of 1 MHz. For such a screen, there are 22 meshes per linear inch; the center-of-wire to center-of-wire distance is 1/22 (0.045) inch and the opening width is smaller by an amount equal to the wire diameter, 0.015 inches. The depth of the apertures is assumed equal to the wire diameter.

Thus

$$A_a = (27.3)D/W = (27.3)(.015) / (0.045 - 0.015)$$

$$= 13.65 \text{ dB}$$

The impedance ratio for the magnetic wave and rectangular apertures is given by

$$k = W/\pi r = (0.045 - 0.015) / \pi(1.75)$$

$$= 0.00546$$

and the reflection term is

$$R_a = 20 \log \left| \frac{(1+k)^2}{4k} \right| = 33.3 \text{ dB}$$

The multi-reflection correction term is

$$C_a = 20 \log \left| 1 - \frac{(k-1)^2}{(k+1)^2} \times 10^{-A_a/10} \right|$$

$$= -0.4 \text{ dB.}$$

The number of openings correction factor is

$$K_1 = 10 \log \frac{1}{an}$$

$$= 10 \log \frac{1}{(0.045 - 0.015)^2 (22)^2}$$

$$= 3.5 \text{ dB.}$$

The skin depth correction factor is

$$K_2 = -20 \log (1 + (35/p)^{2.3})$$

$$p = \frac{0.015}{2.6 \times 10^{-3}} = 5.77$$

$$K_2 = -20 \log (1 + (35/56.3)) = -4.2 \text{ dB.}$$

Finally, the hole-coupling correction factor is given by

$$K_3 = 20 \log (1/\tanh(A_g/8.886))$$

$$= 0.8 \text{ dB.}$$

The screen's shielding effectiveness, SE, is the sum of the six factors:

$$SE = 13.5 + 33.2 - 0.4 + 3.5 - 4.2 + 0.8$$

$$= 46.4 \text{ dB.}$$

Figure 8-18 presents both calculated and measured values of shielding effectiveness for several types of copper screen located 1.75 inches from a loop antenna. Representative non-solid sheet shielding effectiveness measurements are shown in Tables 8-12 and 8-13. The two tables provide data on a variety of material forms, including meshes, perforated sheets, and cellular structures against low-impedance, high-impedance, and plane waves. Figures 8-19 and 8-20 illustrate how the effectiveness of perforated sheet material changes with changes in hole size and hole separation. Table 8-14 contains both calculated and measured values of shielding effectiveness for the No. 22, 15 mil copper screen of the example for magnetic, plane, and electric waves of several frequencies. The shielding effectiveness of the screen is seen to increase with the frequency for magnetic fields, to decrease with increasing frequency for plane waves, and to be largely independent of frequency for electric fields.

Screen shields should use a single or double layer of copper or brass mesh of No. 16 or 22 gauge wire with openings no greater than 1/16 inch. A mesh less than 18 by 18 (wires to the inch) should not be used. The mesh wire diameter should be a minimum of 0.025 inch (No. 22 AWG). If more than a nominal 50 dB of attenuation is required, the screening should have holes no larger than those in a 22 by 22 mesh made of 15 mil of copper wires. The attenuation of an electromagnetic wave by a mesh is considerably less than that afforded by a solid metal screen. The principal shielding action of a mesh is due to reflection. Tests have shown that mesh with 50 percent open area and 60 or more strands per wave length introduces a reflection loss very nearly equal to that of a solid sheet of the same material.

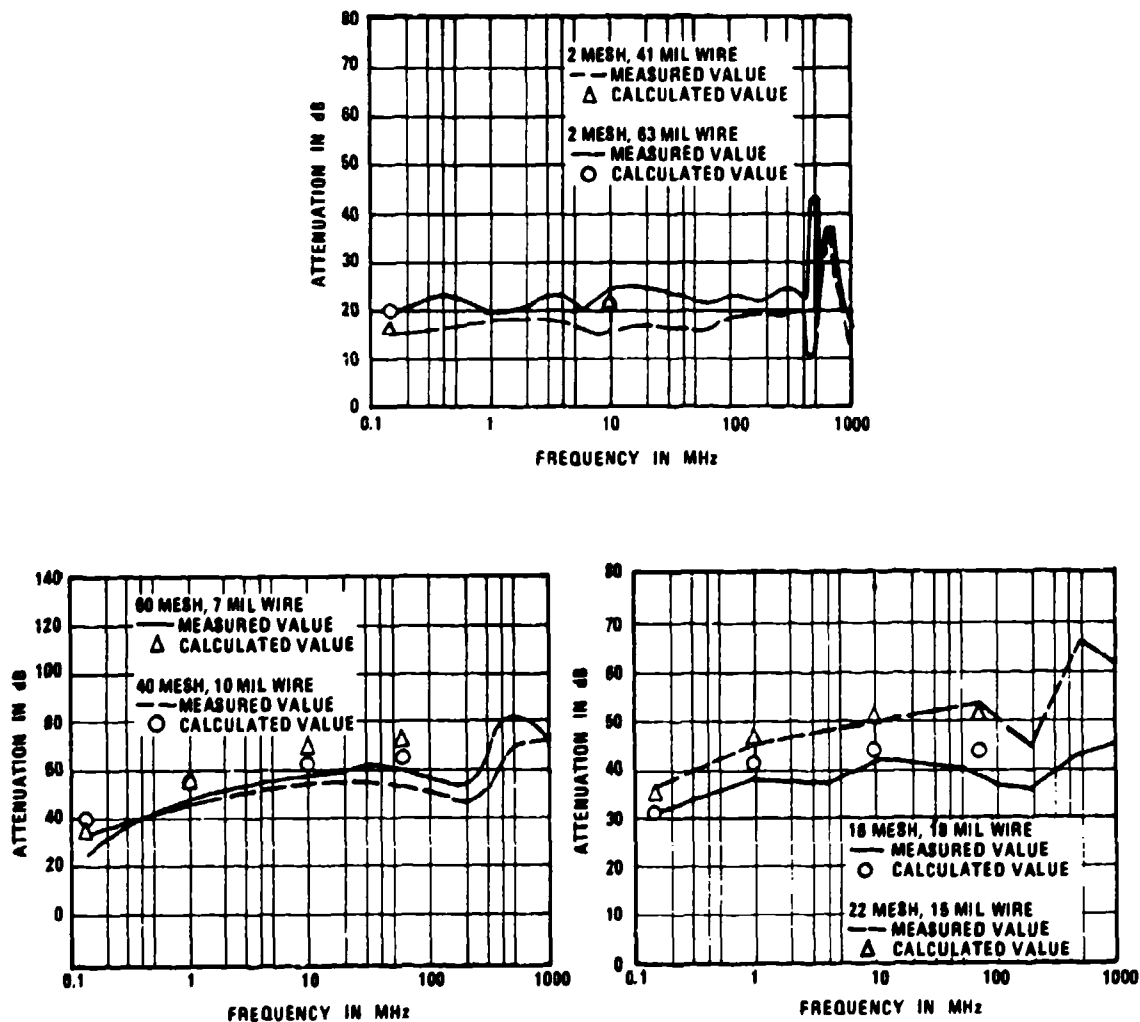


Figure 8-18. Measured and Calculated Shielding Effectiveness of Copper Screens to Low Impedance Fields (8-8)

Table 8-12

Effectiveness of Non-Solid Shielding  
Materials Against Low Impedance and Plane Waves (8-7)

Impinging Wave	Form		Material	Thickness (mils)	Nominal Effectiveness (dB)						
	General	Detail			0.1 kHz	1 kHz	10 kHz	85 kHz	1 MHz	10 MHz	
Low Impedance	Mesh (Screening)	2 layers 1 inch apart	Cu (oxidized)		2	6	18				
		No. 22	Cu				31	43	43		
		No. 16	Bronze				18				
		No. 4	Galvanized Steel				10	17	21		
Plane	Perforated Sheet	45 mil dia., 225 sq. inch	Al	20	3040 MHz			9380 MHz			
					60		62				
	Mesh (Screening)	No. 16	Al	dia. = 13	34		36				
		No. 22	Cu	dia. = 15	118	200 kHz	1 MHz	5 MHz	100 MHz	80	

Table 8-13  
Effectiveness of Non-Solid Shielding Materials Against High Impedance Waves (8-3)

General Form	Detailed	Material	Thickness (mils)	Nominal Effectiveness (14 kHz to 1000 MHz) (dB)	Open Area %	Air-Flow Static Pressure	
						200 cu ft/min (inches of water)	400 cu ft/min
Hexcell	1/4-inch cell, 1 inch thick	Al	3	>90		0.06	.26
TV Shadow Mask (Photo- Etched)	9-mil holes, 28-mil centers	95% Cu 5% 100%	7	>90	12 50 50	>2 0.2 0.2	0.4 0.5
			3	>90			
Lektromesh	40 count	Cu-Ni	7	>90	36	0.4	1.7
	25 count		5	78	49	0.2	0.5
	40 count 25 count	Cu	3	78	57 56	0.2 0.2	0.5 0.4
Perforated Sheet	1/8-inch dia, 3/16-inch	Steel	60	58		0.27	.0.6
	1/4-inch dia, 5/16-inch centers	Al	60	48	46		
	7/16-inch dia, 5/8-inch centers		37	35	45		
Mesh (Screen- ing)	No. 16	Al	20 (dia)	55	36		
	16 x 16/sq in. No. 22	Cu		65 (14 kHz to 60 MHz) 50	50		
	No. 12		20 (dia)	45 (14 kHz to 60 MHz) 40			
	No. 16	Bronze		35	76		
	No. 10	Monel	18 (dia)				
	No. 4	Galva- nized Steel	30 (dia)	28 (14 kHz to 40 MHz) 24			
	No. 2				88		



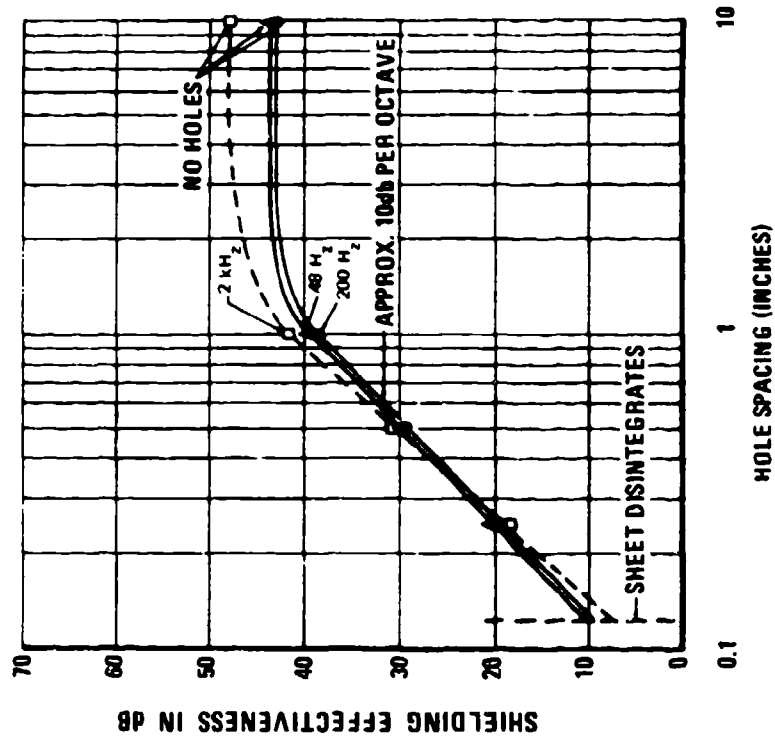


Figure 8-20. Shielding Effectiveness of a Perforated Metal Sheet as a Function of Hole Spacing (8-6)

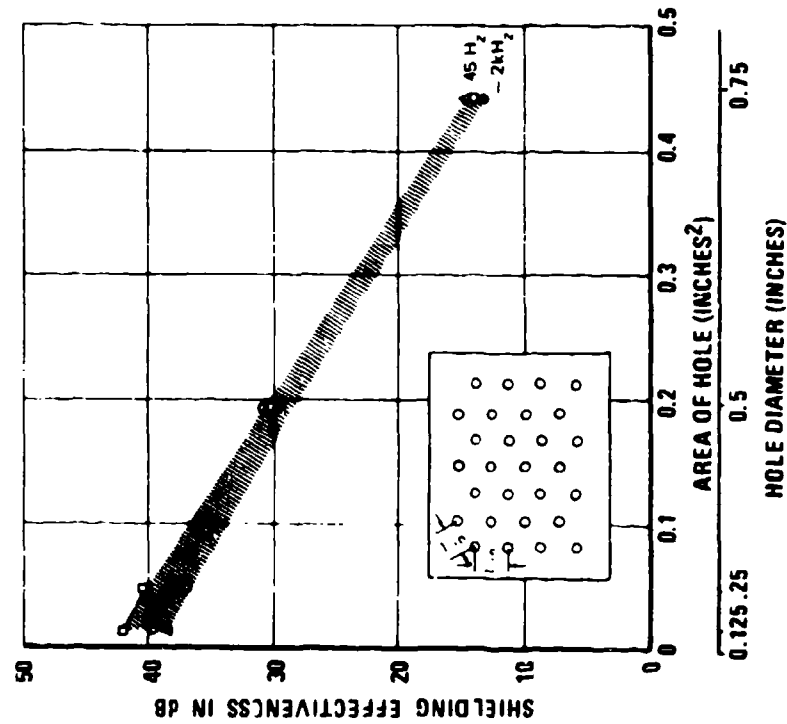


Figure 8-19. Shielding Effectiveness of a Perforated Metal Sheet as a Function of Hole Size (8-6)

Table 8-14

Comparison of Measured and Calculated Values  
of Shielding Effectiveness for No. 22, 15 Mil Copper Screens (8-8)

<u>Test Type</u>	<u>Frequency (MHz)</u>	<u>Measured Effectiveness (dB)</u>	<u>Calculated Effectiveness (dB)</u>
Magnetic field ( $r = 1.75''$ )	0.085	31	29
	1.000	43	46
	10.000	43	49
Plane Wave	0.200	118	124
	1.000	106	110
	5.000	100	95
	100.000	80	70
Electric field	0.014	65	* 65

\*\*The value assumes a wave impedance equal to that of a 30-inch square waveguide.

The mesh construction should have individual strands permanently joined at points of intersection by a fusing process so that a permanent electrical contact is made and oxidation does not reduce shielding effectiveness. A screen of this construction will be very effective for shielding against electric (high-impedance) fields at low frequencies because the losses will be primarily caused by reflection. Installation can be made by connecting a screen around the periphery of an opening.

#### 8.5 SHIELD DISCONTINUITY EFFECTS (APERTURES).

An ideal shielded enclosure would be one of seamless construction with no openings or discontinuities. However, personnel, powerlines, control cables, and/or ventilation ducts must have access to any practical enclosure. The design and construction of these discontinuities become very critical in order to incorporate them without appreciably reducing the shielding effectiveness of the enclosures. Since most mechanically suitable metal enclosures will give enough shielding above 1 MHz, EMI leakage above 1 MHz is due primarily to discontinuities. EMI leakage (the amount of EM energy that will leak from a discontinuity) depends mainly on:

- a. maximum length (not area) of the opening,

- b. the wave impedance, and
- c. the wavelength of the EM energy.

Maximum length rather than width of an opening is important because the voltage will be highest wherever the "detour" for the currents is longest. This is at the center of the slot and the voltage increases as the length of the slot increases. The width has almost no effect on "detour" length and as a consequence has little effect on the voltage.

Wavelength controls how much the "slot antenna" radiates. If the slot happens to be  $1/4$  wavelength or longer, it will be a very efficient radiator; if it is less than  $1/100$  wavelength, it will be a rather inefficient radiator. Therefore, slots only .001" to .005" wide but  $1/100$  wavelength or more long can be responsible for large leaks. Figure 8-21 shows wavelength and  $1/100$  wavelength vs frequency for 0"-6" slot lengths typical in normal metal enclosures. Combinations of frequency and slot lengths to the right of the  $1/100$  wavelength line would tend to be leaky. This figure shows why discontinuities in shields, even if very narrow but a few inches long, will severely reduce the shielding capacity of an enclosure above 100 MHz.

Some types of discontinuities commonly encountered include:

- a. Seams between two metal surfaces, with the surfaces in intimate contact (such as two sheets of material that are riveted or screwed together),
- b. Seams or openings between two metal surfaces that may be joined using a metallic gasket, and
- c. Holes for ventilation or for exit or entry of wire, cable, light, film, water, meter faces, etc.

#### 8.5.1 Seams Without Gaskets.

Seams or openings in enclosure or compartment walls that are properly bonded will provide a low impedance to rf currents flowing across the seam. When good shielding characteristics are to be maintained, permanent mating surfaces of metallic members within an enclosure should be bonded together by welding, brazing, sweating, swagging, or other metal flow processes. To insure adequate and properly implemented bonding techniques, the following recommendations should be observed:

- a. All mating surfaces must be cleaned before bonding.
- b. All protective coatings having a conductivity less than that of the metals being bonded must be removed from the contact areas of the two mating surfaces before the bond connection is made.
- c. When protective coatings are necessary, they should be so designed that they can be easily removed from mating surfaces prior to bonding. Since the mating of bare metal to bare metal is essential for a satisfactory bond, a conflict may arise between the bonding and finish specifications. From the viewpoint of shielding effectiveness, it is preferable to remove the finish where a compromise of the bonding effectiveness would occur.

d. Certain protective metal platings such as cadmium, tin, or silver need not in general be removed. Similarly, low-impedance corrosion-resistant finishes suitable for aluminum alloys, such as alodine, iridite, oakite, turco and bonderrite, may be retained. Most other coatings, such as anodizing, are nonconductive and should be removed. See Figure 8-22 for shielding effectiveness degradation data on selected surface finishes.

e. Mating surfaces should be bonded immediately after protective coatings are removed to avoid oxidation. Refinishing after bonding is acceptable from the standpoint of shielding effectiveness.

f. When two dissimilar metals must be bonded, metals that are close to one another in the electrochemical series should be selected in order to reduce corrosion.

g. Soldering may be used to fill the resulting seam, but should not be employed to provide bond strength.

h. The most desirable bond is achieved through a continuous butt or lap weld. Spot welding is less desirable because of the tendency for buckling, and the possibility of corrosion occurring between welds. Riveting or pinning is even less desirable because of the greater susceptibility of bond degradation with wear.

i. An overlap seam, accompanied by soldering or spot welding, provides a relatively effective bond. Other types of crimped seams may be employed so long as the crimping pressure is uniformly maintained.

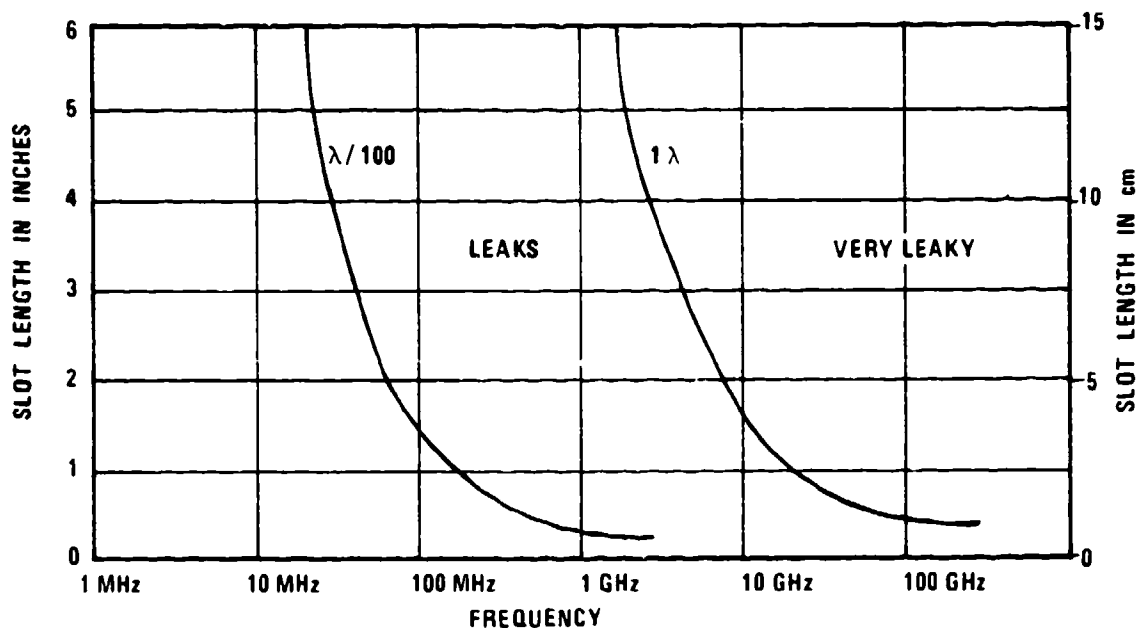


Figure 8-21. Slot Radiation (Leakage) (8-9)

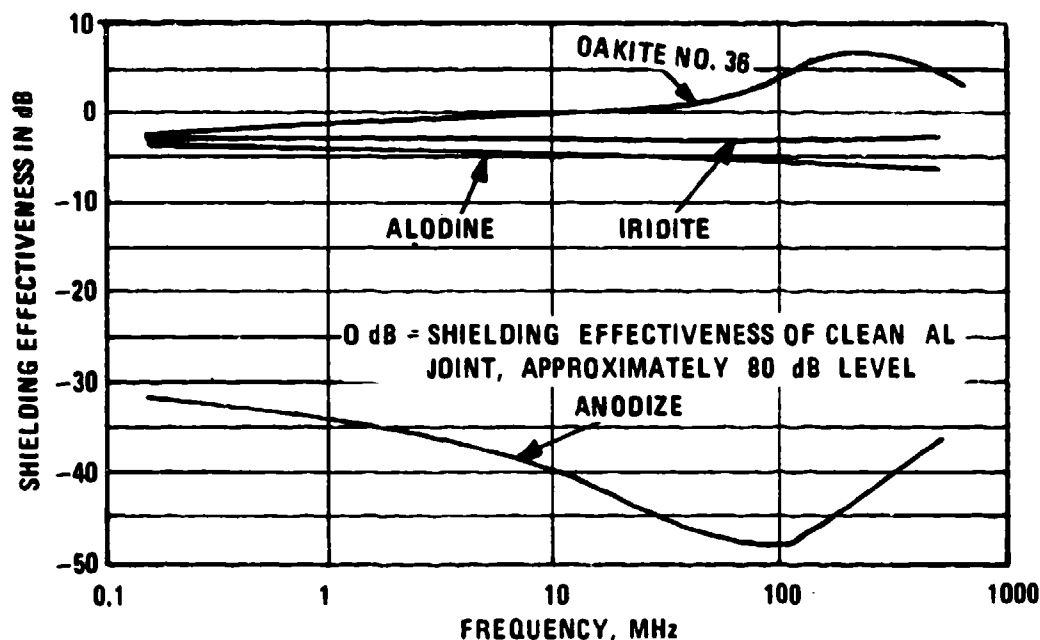


Figure 8-22. Shielding Effectiveness Degradation Caused by Surface Finishes on Aluminum (8-4)

There are often occasions when good temporary bonds must be obtained. Bolts, screws, or various types of clamp and slide fasteners have been used for this purpose. The same general requirements of clean and intimate contact of mating surfaces, and minimized electrolytic (cathodic) effects apply to temporary bonds as well. Positive locking mechanisms that ensure consistent contact pressure over an extended period of time should be used.

Bolts, nuts, screws, and washers that must be manufactured with material different from the surfaces to be bonded should be higher in the electromotive series than the surfaces themselves so that any material migration erodes replaceable components.

A critical factor in temporary bonds (and in spot-welded permanent bonds as well) is the linear spacing of the fasteners or spot welds. Figure 8-23 provides an indication of the sensitivity of this parameter for a 1.27 cm (1/2-inch) aluminum lap joint at 200 MHz. The shielding effectiveness shown in 2.54 cm (1-inch) spacing is about 12 dB poorer than an identical configuration incorporating a 1.27 cm (1/2-inch) wide monel mesh gasket; the effectiveness at 25.4 cm (10-inch) spacing is about 30 dB poorer than that with the same gasket. Use of conductive gaskets for this and other applications is discussed in the next section.

Similar techniques to those just described can be employed in connection with seams in magnetic materials. Permanent seams can be butt or lap, continuous or spot welded using an electric arc in an argon or helium

atmosphere, recognizing that a final material heat treatment will be necessary. Temporary seams are usually screwed or bolted together. Figures 8-24 and 8-25 indicate the change in shielding effectiveness of an AMPB-65 seam at various frequencies as a function of screw spacing and lap joint width, respectively.

#### 8.5.2 Seams With Gaskets.

Considerable shielding improvement over direct metal-to-metal mating of shields used as temporary bonds can be obtained using flexible, resilient metallic gaskets placed between shielding surfaces to be joined. Clean metal-to-metal mating surfaces and a good pressure contact are necessary.

The major material requirements for rf gaskets include compatibility with the mating surfaces, corrosion resistance, appropriate electrical properties, resilience (particularly when repeated compression and decompression of the gasket is expected), mechanical wear, and ability to form into the desired shape. On this basis, monel and silver-plated brass are generally the preferred materials, with aluminum used only for gasketing between two aluminum surfaces. Beryllium-copper contact fingers are also employed, with a variety of platings available, if desired. Mumetal and Permalloy have been used when magnetic shielding effectiveness is of concern.

Gaskets are manufactured with rubber or neoprene to provide both fluid and conductive seals, or to sustain a pressure differential, as well as provide an rf barrier. They are also made using sponge silicon for high temperature applications and are made with both nonconductive or conductive pressure sensitive adhesives. A few of the gasket design approaches that have been employed are summarized in Table 8-15. Typical gasket mounting techniques are given in Figure 8-6. The most frequently used gasket configuration is the knitted wire mesh; the structure of this mesh is shown in Figure 8-27.

The necessary gasket thickness is dependent on the unevenness of the joint to be sealed, the compressibility of the gasket, and the force available. The shape required depends on the particular application involved, as well as the space available, the manner in which the gasket is held in place, and the same parameters that influence gasket thickness. Gaskets may be held in place by sidewall friction, by soldering, by adhesives, or by positioning in a slot or on a shoulder. Soldering must be controlled carefully to prevent its soaking into the gasket and destroying gasket resiliency. Adhesives (particularly nonconductive adhesives) should not be applied to gasket surfaces that mate for rf shielding purposes; auxiliary tabs should be used. A recommended pressure is about 20 psi.

**8.5.3 Penetration Holes.** One effective method of neutralizing the shielding discontinuities created by planned holes (e.g., for air ventilation and circuit adjustment) in a shield is to use cylindrical and rectangular waveguide-below-cutoff slots or tubes.

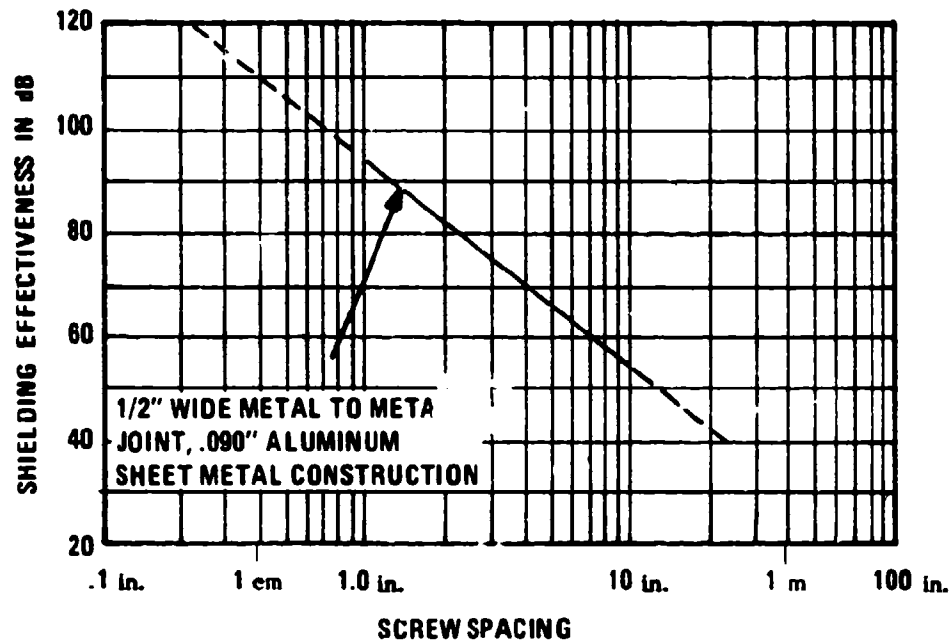


Figure 8-23. Influence of Screw Spacing on Shielding Effectiveness

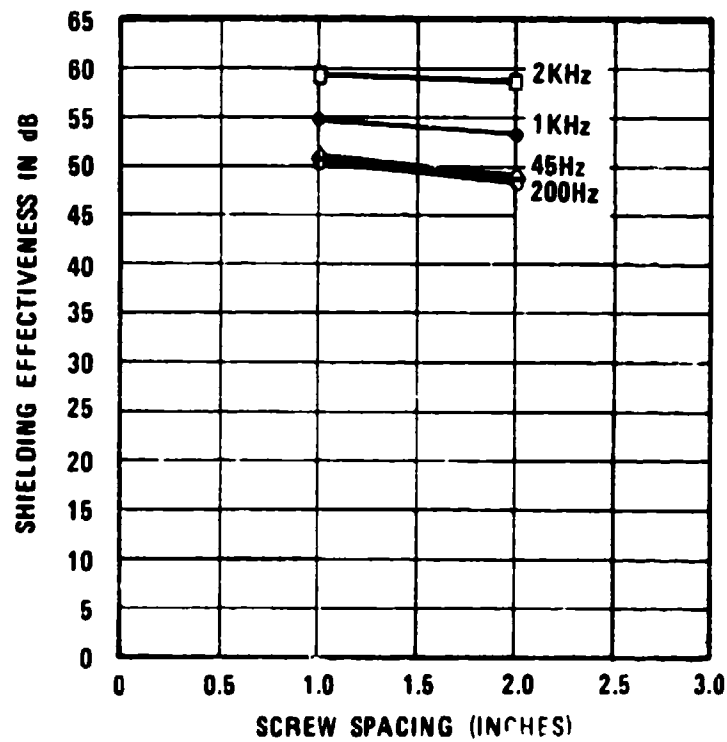


Figure 8-24. Shielding Effectiveness Along the Edge of the Overlap as a Function of Screw Spacing Along the Edge of the Overlap (8-6)

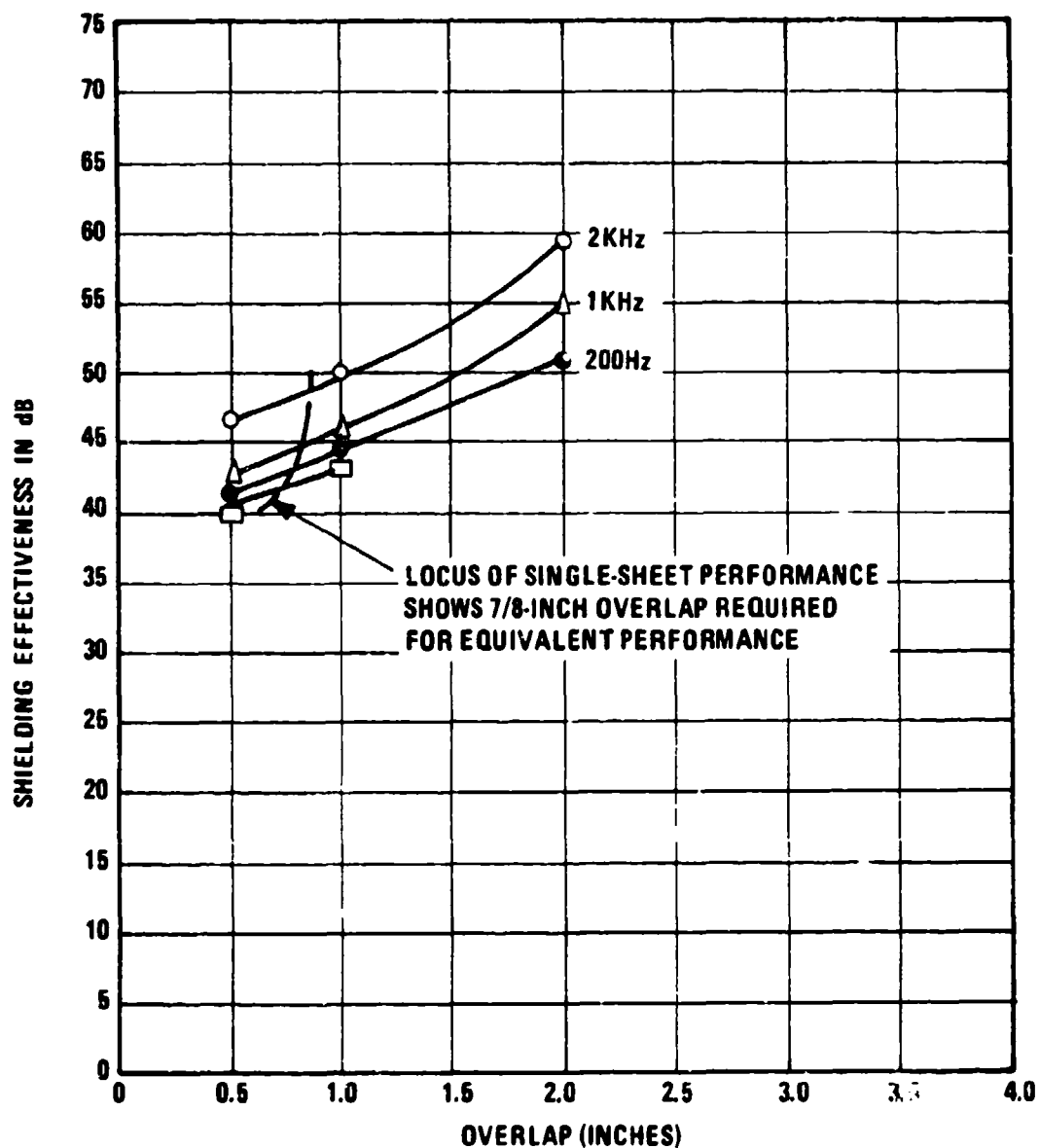


Figure 8-25. Shielding Effectiveness of an AMPB-65 Joint as a Function of Overlap (8-1)



Table 8-15

## Characteristics of Conductive Gasketing Materials

<u>Material</u>	<u>Chief Advantages</u>	<u>Chief Limitations</u>
Compressed knitted wire	Most resilient all-metal gasket (low flange pressure required). Most points of contact. Available in variety of thicknesses and resiliencies, and in combination with neoprene and silicon.	Not available in sheet (Certain intricate shapes difficult to make). Must be 0.40 inch or thicker.
Brass or beryllium copper with punctured holes	Best break-thru of corrosion protection films.	Not truly resilient nor generally reusable.
Oriented wires in rubber or silicon	Combines fluid and rf seal. Can be effective against corrosion films if ends of wires are sharp.	May result in larger size gasket for same effectiveness.
Aluminum screen impregnated with neoprene	Combines fluid and conductive seal. Thinnest gasket. Can be cut to intricate shapes.	Very low resiliency (high flange pressure required).
Soft Metals	Cheapest in small sizes.	Cold flows, low resiliency.
Metal over rubber	Takes advantage of the resiliency of rubber.	Foil cracks or shifts position. Generally low insertion loss yielding poor rf properties.
Conductive rubber	Combines fluid and conductive seal.	Practically no insertion loss, giving very poor rf properties.
Contact Fingers	Best suited for sliding contact.	Easily damaged. Few points of contacts.

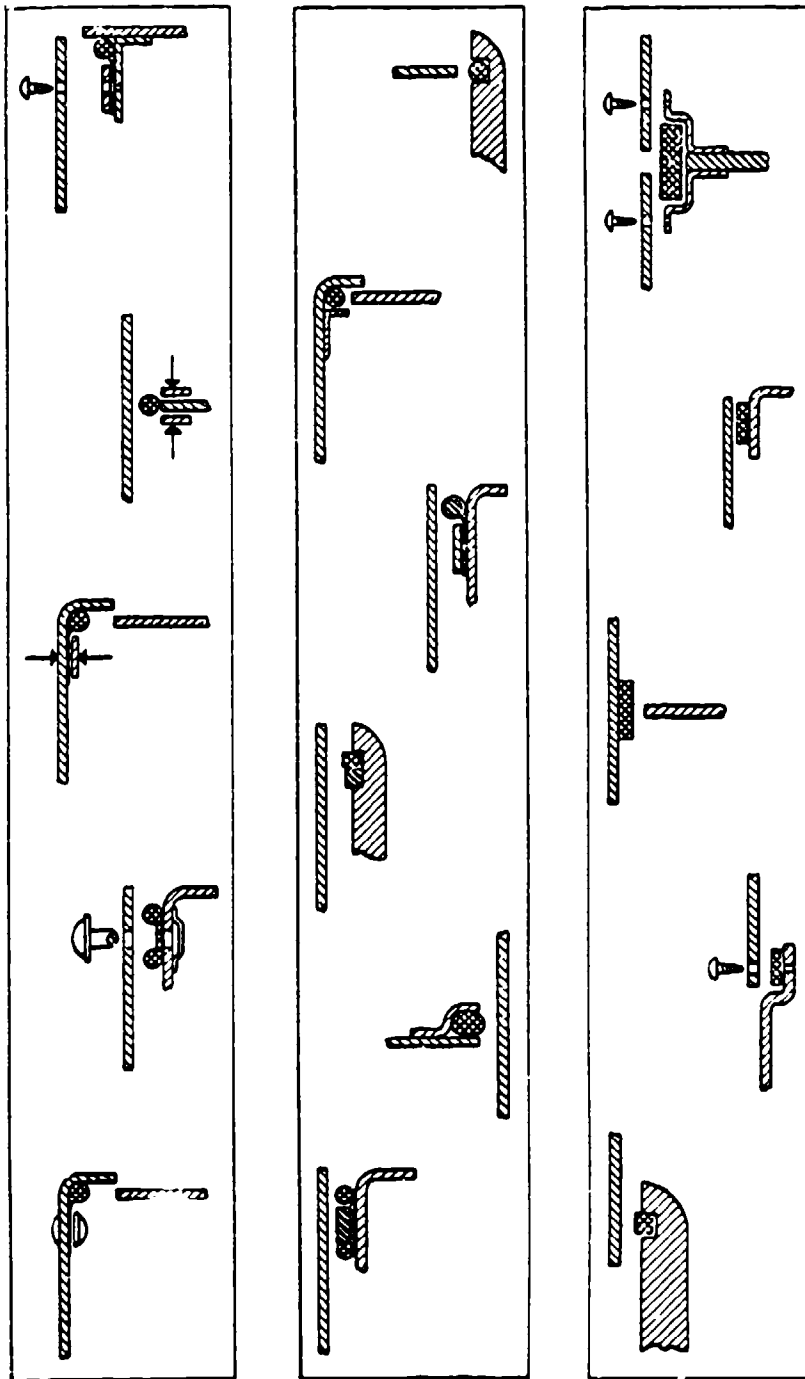


Figure 8-26. Typical Mounting Techniques for RF Gaskets

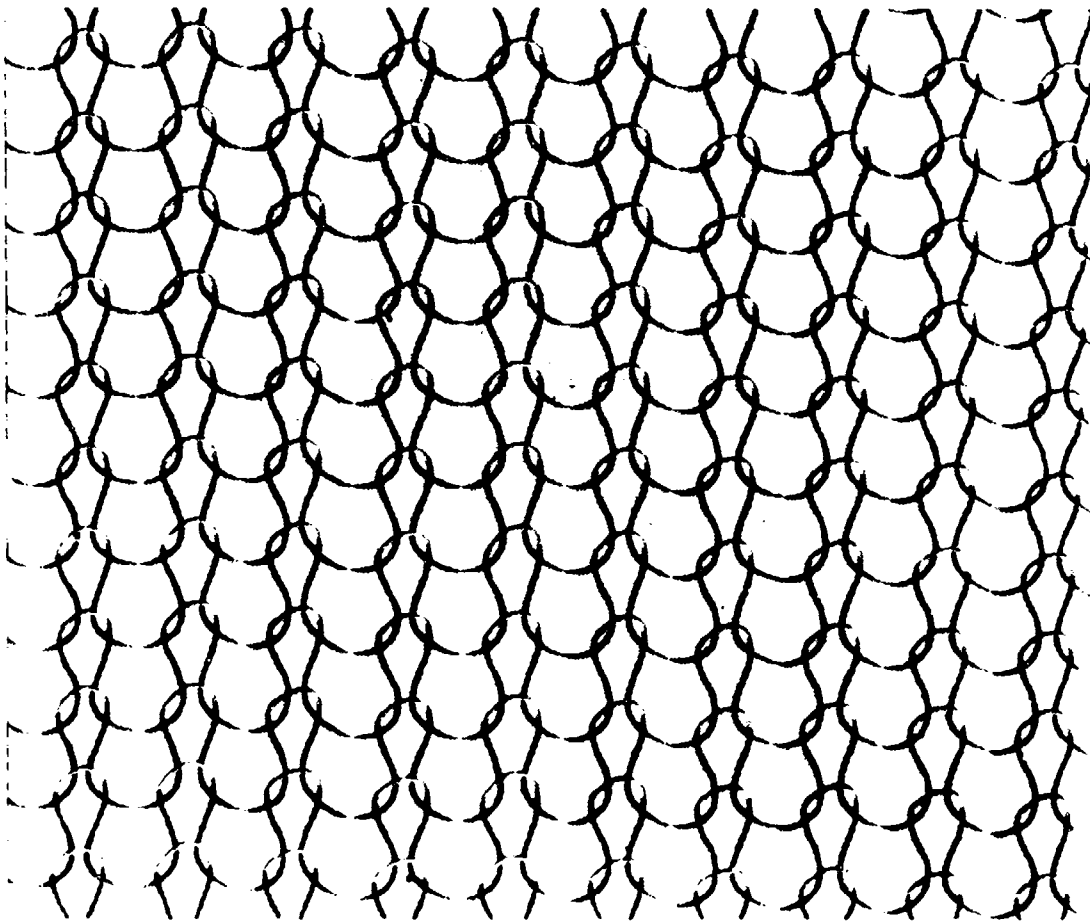


Figure 8-27. Enlarged View of Knitted Wire Mesh

#### 8.5.3.1 Waveguide-Below-Cutoff.

A properly designed waveguide-below-cutoff opening will act like a high-pass filter. The cutoff frequency is a function of the cross-section of the waveguide. For a cylindrical waveguide, the cutoff frequency of the dominant TE mode is

$$f_c = \frac{6.92}{d} \text{ GHz.} \quad (8-25)$$

The cutoff frequency for the TE mode of rectangular waveguide is

$$f_c = \frac{5.90}{b} \text{ GHz.} \quad (8-26)$$

In these equations,

- $f_c$  = cutoff frequency for the dominant mode in gigahertz,
- $d$  = inside diameter of a cylindrical waveguide in inches, and
- $b$  = greatest dimension of rectangular waveguide in inches.

At any frequency,  $f_a$ , considerably less than cutoff (i.e.,  $f_a < 0.1f_c$ ), the attenuation,  $\alpha$ , in dB per inch for cylindrical waveguides is approximated by the relation

$$\alpha = \frac{32}{d} \quad (8-27)$$

For rectangular waveguides, the attenuation,  $\alpha$ , in dB per inch is

$$\alpha = \frac{27.3}{b} \quad (8-28)$$

The equations given above are valid for air-filled waveguides with length-to-width or length-to-diameter ratios of 3 or more.

In many cases, shielding screens introduce excessive air resistance (See Vol II) and may provide inadequate shielding effectiveness. In such cases, openings may be covered with specially designed ventilation panels (such as honeycomb) with openings that operate on the waveguide-below-cutoff principle. The shielding effectiveness of honeycomb panels is a function of the size and length of the waveguide and the number of waveguides in the panel. Table 8-16 indicates the shielding effectiveness of a honeycomb panel constructed of steel with 1/8-inch hexagonal openings 1/2-inch long.

Table 8-16

**Shielding Effectiveness of Hexagonal Honeycomb Made of Steel  
with 1/8-inch Openings 1/2-Inch Long (8-10)**

<u>Frequency</u> (MHz)	<u>Shielding Effectiveness</u> (dB)
0.1	45
50.0	51
100.0	57
500.0	56
2,200.0	47

Honeycomb-type ventilation panels in place of screening:

- a. allow higher attenuation that can be obtained with mesh screening over a specified frequency range,

- b. allow more air to flow with less pressure drop for the same diameter opening,
- c. cannot be damaged as easily as the mesh screen and are therefore more reliable, and
- d. are less subject to deterioration by oxidation and exposure.

All non-solid shielding materials, such as perforated metal, fine mesh copper screening, and metal honeycomb, present an impedance to air flow. Metal honeycomb is the best of these materials because it enables very high electric field attenuations to be obtained through the microwave band with negligible drops in air pressure (see Volume II). However, honeycomb has the disadvantages of occupying greater volume and costing more than screening or perforated metal. Further, it is often difficult to install honeycomb paneling because flush mounting is required. Thus, screening and perforated sheet stock sometimes find application for purely physical design reasons, although honeycomb panels can achieve attenuations greater than 100 dB for frequencies below 10 MHz.

The waveguide attenuator is also of considerable value where control shafts must extend through an enclosure. By making use of an insulated control shaft passing through the waveguide attenuator, the control function can be accomplished with little likelihood of radiation. However, where a metallic control shaft is required, it must be grounded to the case by a close-fitting gasket or metallic fingers.

Fuseholders, phone jacks, panel connectors not in use, and other receptacles can be fitted with a metallic cap that provides an electrically continuous cover and maintains case integrity.

The waveguide attenuator approach may also be considered where holes must be drilled in the enclosure. If the metal thickness is sufficient to provide a "tunnel" with adequate length, a waveguide attenuator is effectively produced. For example, a metal wall 0.5 cm (3/16-inch) thick would permit a 0.16 cm (1/16-inch) hole to be used without excessive leakage. This technique definitely should be considered where it is necessary to confine extremely intense interference sources.

#### 8.5.3.2 Screen and Conducting Glass.

Often it is necessary to provide rf shielding over pilot lights, meter faces, strip chart recorders, oscilloscopes, or similar devices that must be observed by the equipment user. The alternatives available include:

- a. Use of a waveguide attenuator,
- b. Use of screening material,
- c. Providing a shield behind the assembly of concern, and filtering all leads to the assembly, or
- d. Use of conducting glass.

A waveguide attenuator is a practical approach for rf shielding of lamps. The technique has the advantage of not introducing light transmission loss. However, it is not particularly suitable for most meter openings or larger apertures because of the space requirements involved.

Use of screens over meter faces and other large apertures has often been employed for shielding purposes. A typical screen introduces a minimum of 15%-20% optical loss which can create difficulties in reading meters. If the device being shielded has a scale (such as an oscilloscope graticule), bothersome zoning patterns can result. However, these potential deficiencies are counterbalanced by good shielding efficiencies at a fairly low cost.

Glass coated with conducting material such as silver can provide shielding across viewing surfaces with some loss in light transmission. Conductive glass is commercially available from a number of glass manufacturers.

Figure 8-28 provides shielding effectiveness data on 50 and 200 ohms per square silver-impregnated glass against electric arc discharges. Figure 8-29 indicates shielding effectiveness as a function of surface resistance for plane waves in the frequency range from 0.25 to 350 MHz. The light transmission characteristics of this type of glass as a function of surface resistance is presented in Figure 8-30. For effective shielding, good contact to the conducting surface of the glass must be maintained around its periphery.

### 8.6 SELECTION OF SHIELDING MATERIALS.

The selection of the material should be based on its ability to drain off induced electrical charges and to carry sufficient out-of-phase currents to cancel the effects of the interfering field. The inherent characteristics of the metal to consider are its relative conductivity,  $g_r$ , and its relative permeability,  $\mu_r$ . The thickness of the shield and the frequency of the signal to be attenuated are also important.

The selection of proper materials for shielding should be made in accordance with the following basic rules

- a. At low frequencies (LF), only magnetic materials can furnish appreciable shielding against magnetic fields.
- b. For a given material, magnetic fields require a greater shield thickness than do electric fields.
- c. At higher frequencies, smaller shield thickness is required for a given material.
- d. At sufficiently high frequencies, nonferrous materials such as copper and aluminum will give adequate shielding for either electric or magnetic fields.
- e. The electric field component for frequencies from 60 to 800 Hz (i.e., ac power) can readily be shielded with thin sheets of conducting materials such as iron, copper, aluminum, and brass.

For a detailed description of the procedure for selecting a shield material for a facility, see Volume II. Care must be used when adding a shield to a subsystem. For example, a shield placed too close to a circuit in which the circuit Q is a critical factor can cause degradation of performance because the losses in the shield will appear as an effective resistance in the critical circuit, thereby lowering the circuit Q.

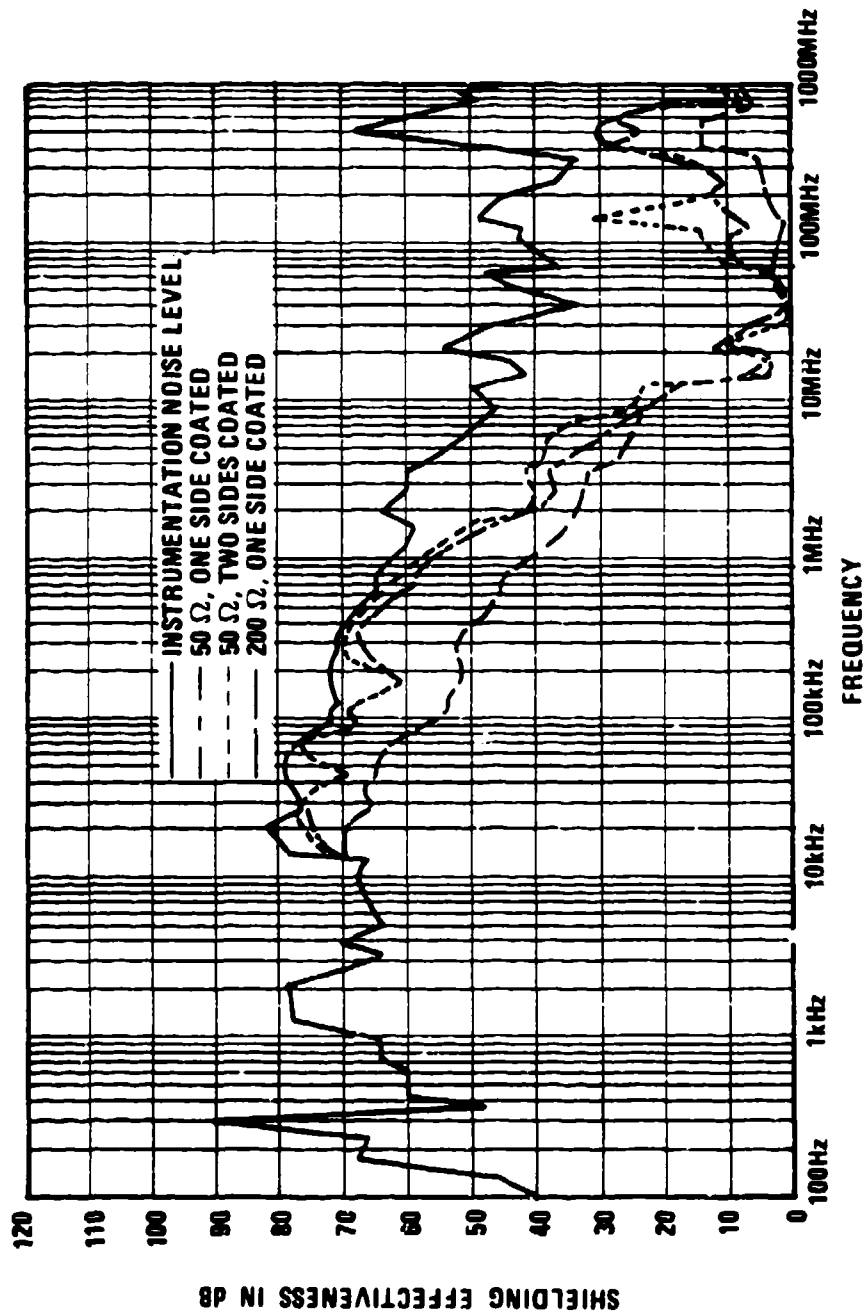


Figure 8-28. Shielding Effectiveness of Conductive Glass to High Impedance Waves (8-11)

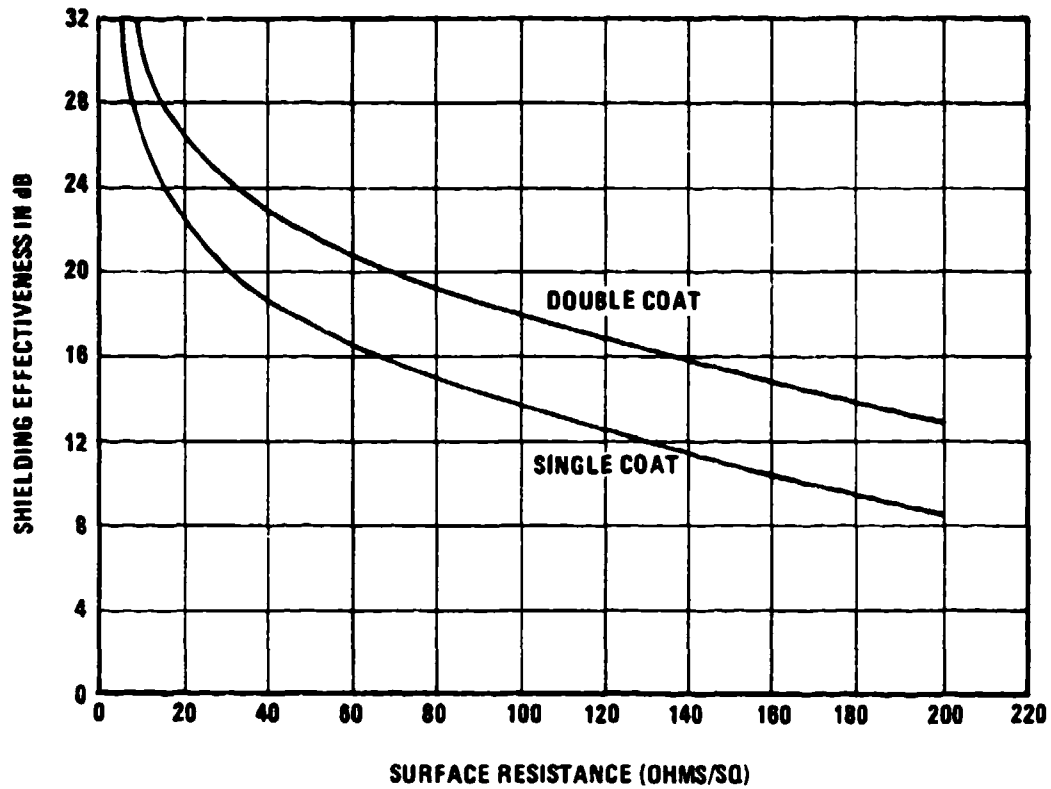


Figure 8-29. Shielding Effectiveness of Conductive Glass to Plane Waves (8-11)

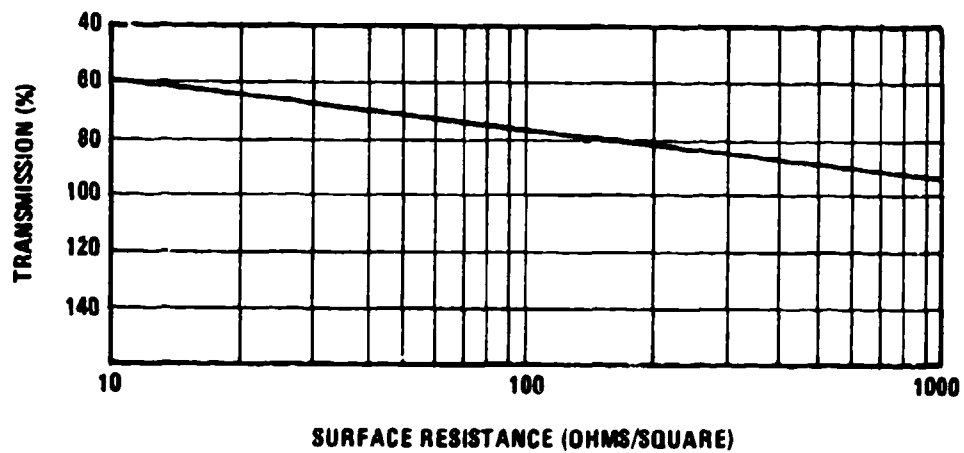


Figure 8-30. Light Transmission Versus Surface Resistance for Conductive Glass (8-7)



**8.7 USE OF CONVENTIONAL BUILDING MATERIALS.** Conventional building materials are not normally selected on the basis of their electromagnetic shielding properties however most materials do provide some limited degree of shielding. Some documented evidence of the shielding provided by common construction materials is available (8-12). Though the data is sketchy, enough does exist to give a preliminary indication of what can be expected from a building made of various materials.

**8.7.1 Concrete.** Figure 8-31 shows that the shielding effectiveness of ordinary concrete is very low. (It may be assumed that the properties of brick are similar to concrete.) The addition of coke and other forms of carbon to concrete can greatly enhance shielding properties. Approximately 30 dB shielding effectiveness from 1 GHz to 10 GHz can be achieved by using concrete and carbon. A concrete-coke aggregate apparently can provide shielding in excess of 30 dB above about 20 MHz and can offer more than 100 dB above 300 MHz.

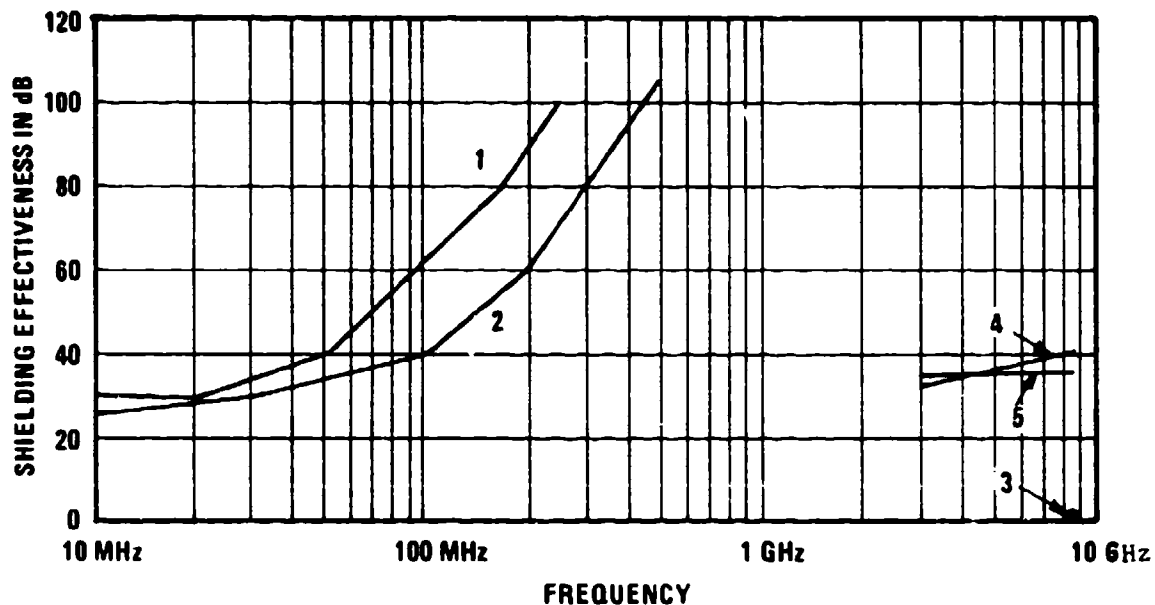
**8.7.2 Reinforcing Steel (Rebar).**

Limited shielding to low frequency fields can be provided by the reinforcing steel or wire mesh in concrete. For maximum shielding, the conductors must be welded at all joints and intersections to form many continuous conducting loops or paths around the volume to be shielded. The degree of shielding will depend on the following parameters:

- a. The size and shape of the volume to be shielded.
- b. The diameter of the bars and spacing (the distance between bar centers).
- c. The electrical and magnetic characteristics of the reinforcement steel materials (conductivity and relative permeability).
- d. The frequency of the incident wave.

The family of curves shown in Figure 8-32 describes the attenuation at approximately 10 kHz for an enclosure whose height is 4.5 meters (15 feet), and other dimensions vary over a 5 to 1 range. Bar diameters are 4.30 cm (1.692 inches) with a spacing of 35.56 cm (14 inches) on centers. The room dimensions, bar spacing, and diameters shown in Figure 8-32 are typical and cover most situations encountered in practice. The values of attenuation indicated are those obtainable at the center of the room. There will be less shielding near the edges of the room. For more detailed design information on the use of reinforcing bars as shields, consult Reference 8-13.

Welded wire fabric imbedded in the walls of a room or building can provide effective shielding if the individual wires of the fabric are joined to form a continuous electrical loop around the perimeter of the area to be shielded. At each seam where the mesh meets, each wire must be welded or brazed to the corresponding wire, or the meshes may be connected by a continuous strap. Additional attenuation may be obtained by use of a double layer of welded wire fabric separated by the thickness of a regular wall.



LEGEND			
CURVE NO.	PHYSICAL CONFIGURATION	TYPE	THICKNESS (MILS)
1	CYLINDER 6' DIA. x 6' LG.	CONCRETE COKE AGGREGATE (NEW)	--
2	" " "	(2 MONTHS OLD)	--
3	---	PLAIN CONCRETE	1000
4	---	CONCRETE, 5% CARBON	1000
5	BLOCK 16' x 8' x 8'	CONCRETE, IMPREGNATED w/CARBON LATEX	

Figure 8-31. Shielding Effectiveness of Some Building Materials (8-12)

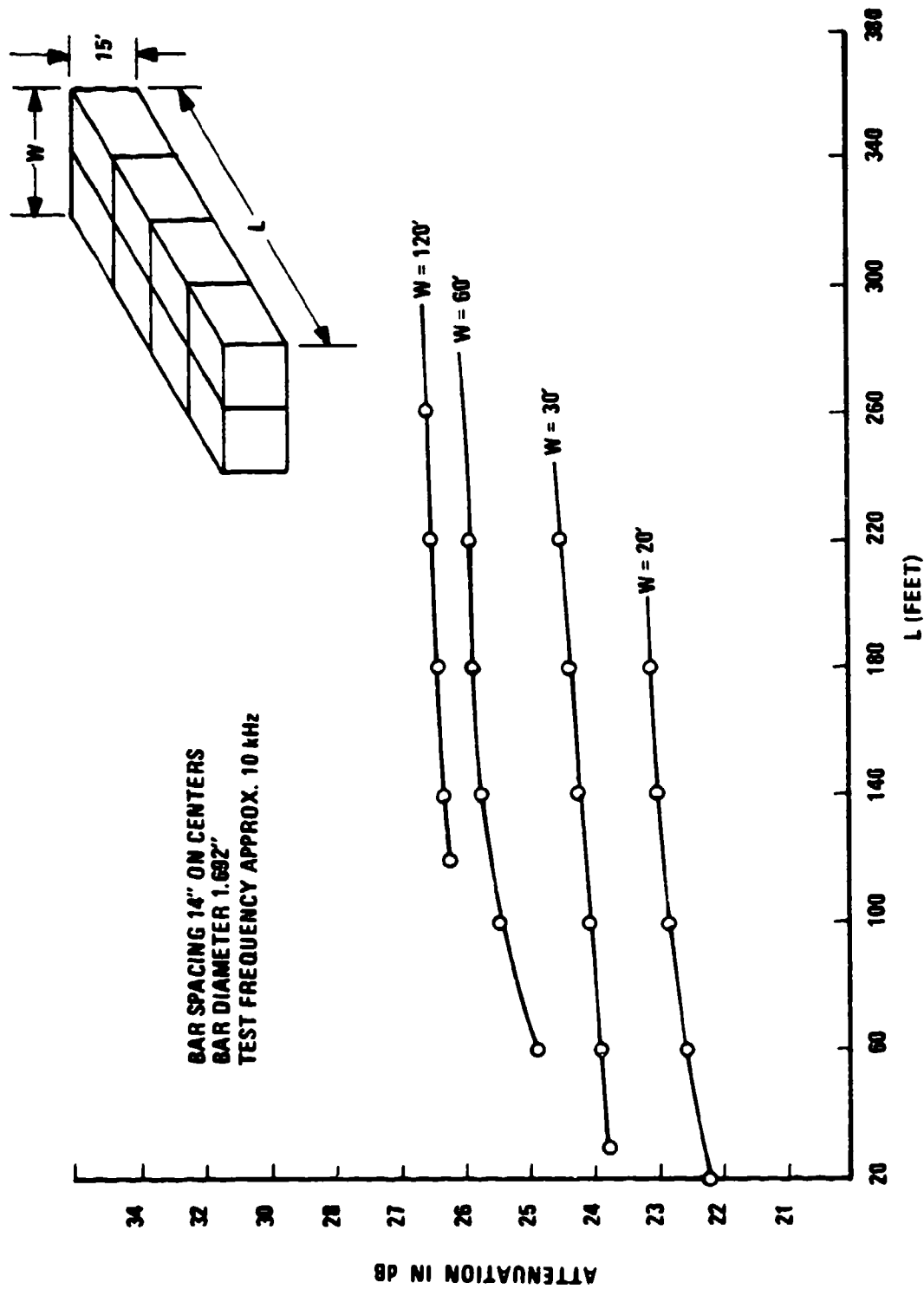


Figure 8-32. Center Area Attenuation of Induced Voltage by 15 Foot High Single-Course Reinforcing Steel Room (8-13)

**8.8 CABLE AND CONNECTOR SHIELDING.** Electromagnetic shielding is required not only for equipment containers but also for many of the cables which connect the equipment units since interference may be transferred from one circuit or location to another by interconnecting cable. The interference may be radiated from a cable or transferred into a cable from external fields. Once interference has been transferred by radiation or common-impedance circuit elements into a cable circuit of an electronic or electrical complex, it can be conducted through interconnecting cables to the other elements of the complex. Because of cable proximity in cable runs or elsewhere, intra- and/or inter-cable crosstalk may occur as a result of electromagnetic transference between cables.

#### 8.8.1 Cable Shields.

The effectiveness of a cable shield is a function of two basic interference mechanisms: (1) EM wave shielding effectiveness and (2) surface transfer impedance,  $Z_t$ . As with other shields, the EM wave shielding effectiveness results from attenuation and reflections and is dependent upon such factors as the type and thickness of the material used and the number and size of openings in the shield. In addition, cable shields frequently are connected in such a manner as to carry relatively large currents themselves. Although the interfering currents generally flow on the outer surfaces of the shields (skin depth effects), an electric field and resulting axial voltage gradient is developed along the inner (shielded) conductor (see Figure 8-33). The ratio of the induced conductor-to-shield voltage per unit length to the shield current is defined as the surface transfer impedance,  $Z_t$ .

The effectiveness of a shield is a function of the conductivity of the metal, contact resistance between strands in the braid, angle and type of weave, strand sizes, percentage of coverage, and size of openings. Analytical expressions which define  $Z_t$  in terms of these parameters are available (8-14). For uniform current distribution along a cable shield, the resulting  $Z_t$  can be used to predict the shield effectiveness of the cable knowing the terminating impedances of the cable. Typically, the cable is several wavelengths long at the frequency of the impinging field. Thus, the current distribution on the cable sheath varies with length and is a function of its orientation to the incident wave and to the surroundings. Since the current distribution will be essentially unpredictable for other than very specialized conditions, the ability to predict shielding effectiveness of the cable shield through the use of  $Z_t$  is severely limited.

There are several methods for shielding cables. These include: (a) braid, (b) flexible conduit, (c) rigid conduit, and (d) spirally-wound shields of high permeability materials. The principal types of shielded cables that are available include shielded single wire, shielded multi-conductor, shielded twisted pair, and coaxial. Cables are also available in both single and multiple shields in many different forms and with a variety of physical characteristics. The general properties of five classes of cable shields are given in Table 8-17.

Braid, consisting of woven or perforated material, is used for cable shielding in applications where the shield cannot be made of solid material. Advantages are ease of handling in cable makeup and lightness in weight. However, it must be remembered that for radiated fields the shielding effectiveness of woven or braided materials decreases with increasing frequency and increases with the density of the weave (8-14). The relative shielding effectiveness of single and double braided cables as a function of frequency is shown in Figure 8-34.

Table 8-17

## Comparison of Cable Shields

	Single <sup>+</sup> Layer <u>Braid</u>	Multiple <sup>+</sup> Layered <u>Braid</u>	<u>Foil <sup>++</sup></u>	<u>Conduit <sup>++</sup></u>	<u>Flexible Conduit</u>
Shield Effectiveness* (Audio Frequency)	Good	Good	Exc.	Exc.	Good
Shield Effectiveness* (Radio Frequency)	Good	Exc.	Exc.	Exc.	Poor
Normal Coverage	60-95%	95-97%	100%	100%	90-97%
Fatigue Life	Good	Good	Fair	Poor	Fair
Tensile Strength	Exc.	Exc.	Poor	Exc.	Fair

\*Poor < 20 dB; Fair, 20-40 dB; Good, 40-60 dB; Exc. > 60dB.

+Effectiveness against magnetic fields is poor.

++For effective magnetic shield, high permeability material must be used.

Conduit either solid or flexible, or zippered tubing may also be used to shield system cables and wiring from the rf environment. The shielding effectiveness of solid conduit is the same, for rf purposes, as that of a solid sheet of the same thickness and material. Linked armor or flexible conduit may provide effective shielding at lower frequencies, but at higher frequencies the openings between individual links can take on slot-antenna characteristics, seriously degrading the shielding effectiveness. If linked armor conduit is required, all internal wiring should be individually shielded. Degradation of conduit shielding is usually not because of insufficient shielding properties of the conduit material but rather the result of discontinuities in the cable. These discontinuities usually result from poor splicing or from improper termination of the shield. Zippered tubing may provide greater than 60 dB of shielding to frequencies below 1 GHz.

For protection against primarily magnetic fields, shielding materials with high permeability are necessary. For example, iron or steel conduit offers better protection against magnetic fields than does aluminum conduit. In lieu of ferrous conduit, annealed high permeability metal strips wrapped around the cable are sometimes used. Multiple layers of counterspiral-wound nickel-iron or silicon-iron alloys, or low carbon steel frequently prove effective. High permeability tape is also available with or without adhesive backing. Also, combination high permeability, high conductivity tape is available which provides both electric and magnetic shielding.

The proper installation of cables is essential if interference difficulties are to be avoided. Assuming proper grounding techniques have been employed, the following are suggested as guidelines for good signal cable practice.

- a. Choose the cable to be utilized according to the characteristic impedance desired, amount of signal attenuation permitted, environment within which the cable must exist, and characteristics of the signal to be transmitted.
- b. Where a high degree of shielding is needed, cables with multiple shields separated by insulation should be used. Double shielding is not effective unless each shield is insulated from each other.
- c. Overall shields of multipair cables should not be used for signal return paths.
- d. Individually shielded cables, where used, should have insulating sleeves or coverings over the shields. Balanced signal circuits should use twisted pair or a balanced coaxial line with a common shield. A coaxial line with a shield is commonly called a triaxial cable. Where multiconductor twisted pair cables that have individual shields as well as a common shield are used, all shields should be insulated from one another within the cable.
- e. Coaxial cables should, in all cases, be terminated in their characteristic impedances.
- f. Coaxial cables carrying high-level energy signals should not be bundled with unshielded cables or shielded cables carrying low-level signals.
- g. Grounding a number of conductor shields by means of a single wire to a connector ground pin should be avoided, particularly if the shield-to-connector, connector-to-ground lead length exceeds one inch, or where different circuits that may interact are involved. Such a ground lead is a common impedance element across which interference voltages can be developed and transferred from one circuit to another.

Great care should be taken at connectors if impedance characteristics and shielding integrity are to be maintained. A shielding shell should be used to shield the individual pins of a connector; a well-designed connector has a shielding shell enclosing its connecting points. The shell of multiple connectors should be connected to the shield. Coaxial lines should terminate in shielded pins. The use of pigtail connections for coaxial lines is undesirable since it permits rf leakage.

Serious interference problems arise when shielded wires or coaxial cables are not properly terminated at the connector. It is important that the connector be properly grounded. The direct bonds for this ground can be achieved by maintaining clean metal-to-metal contact between the connector and equipment housing. In those cases where a large number of individual shields from shielded wires must be connected to ground, it is recommended that the halo technique be used. The exposed unshielded leads should be as short as physically possible to reduce electrical coupling between conductors. Interference is caused when a shielded cable is run into a completely sealed box, but is grounded internally. The correct way to install a shielded rf cable is to run the shield well inside the connector and bond it around the connector shell.

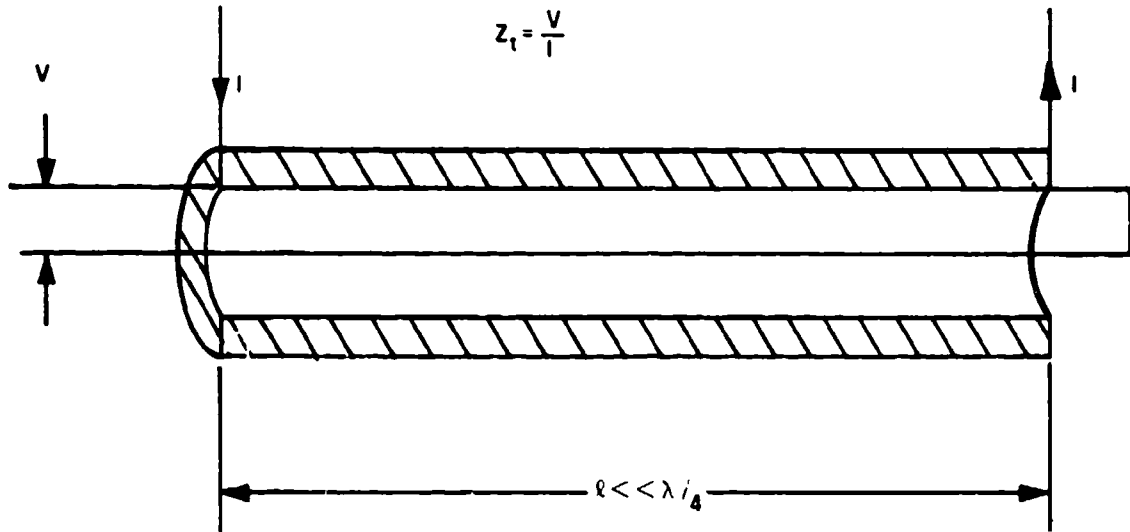


Figure 8-33. Surface Transfer Impedance

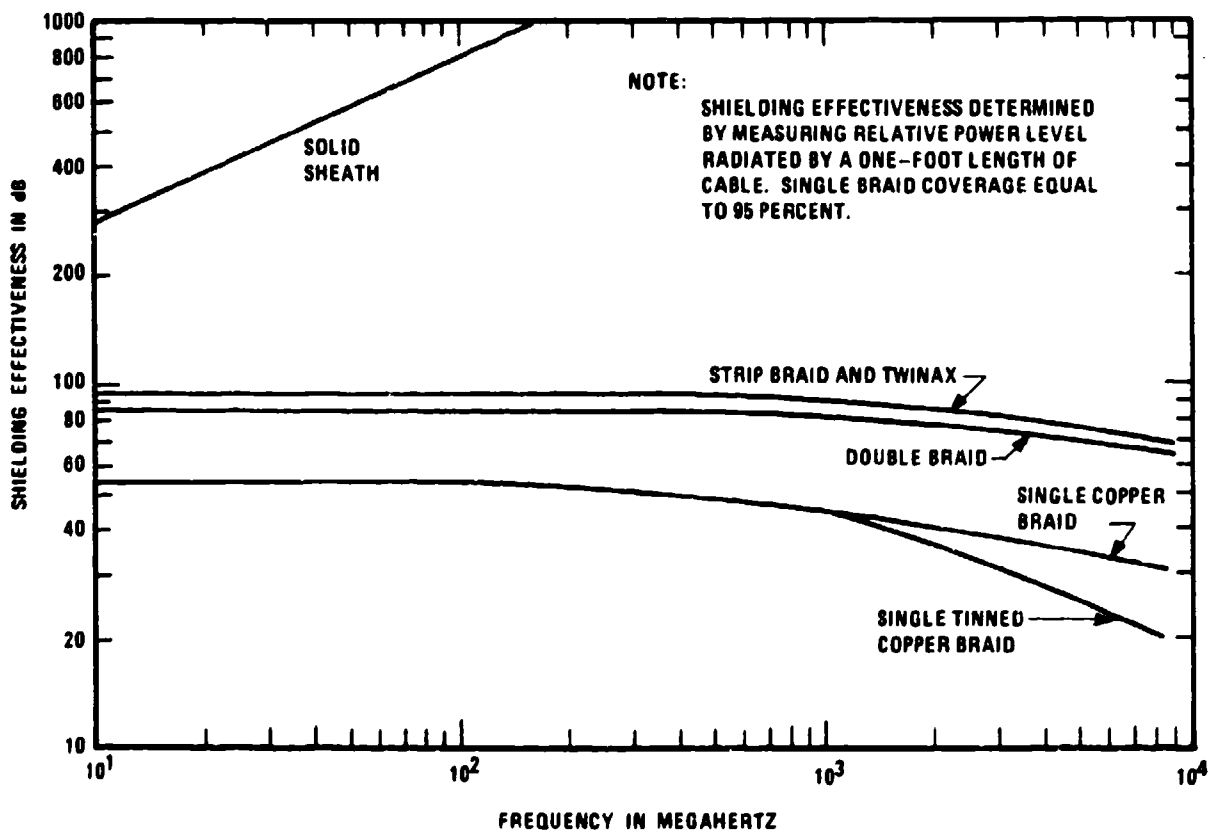


Figure 8-34. Shielding Effectiveness of Various Types of RF Cables as a Function of Frequency (8-13)

### 8.8.2 Terminations and Connectors.

If the effectiveness of a shield is to be maintained, the cable shield must be properly terminated. In an otherwise adequately shielded system, rf currents that are conducted along shields can be coupled to the system wiring from the point of an improper cable termination. This is a particularly important consideration in the case of cables exposed to high power rf fields.

In a properly terminated shield, the entire periphery of the shield is grounded to a low impedance reference, minimizing any potentials at the surface of the termination. MIL-E-45782B (8-16) recommends against use of soldering to terminate shields because of the danger of damaging conductor insulation, and suggests a variety of termination methods, all involving crimping operations. The use of silver epoxy or other synthetic conducting material has been found to be unacceptable for shield bonding because of lack of mechanical strength necessary for this application.

Cable connectors are made in many styles for a multitude of power, signal, control, instrumentation, transducer, audio, video, pulse, and rf applications. They are made to fulfill special functions and may be required to be hermetically sealed, submersion proof, and weatherproof. They are manufactured in the straight type, angle type, screw-on type, bayonet twist-and-lock type, bayonet screw-on-type, barrier type, straight plug-in type, and push-on types (see Table 8-18).

Figure 8-35 illustrates the type of connector that should be used when a shielded cable assembly contains individually shielded wires. The practice of pigtailng these shields and connecting them to one of the pins is not recommended. The individual shields should be connected to coaxial pins specifically adapted for this purpose, with the shields of the mating surfaces making contact before the pins.

When maintaining the shielding integrity of a connector pair (i.e., two interconnecting connectors), a good method to employ (see Figure 8-36) is to place spring contacts inside one portion of one connector so that positive contact is made along the circumference of the mating parts. These contacts are extended so that the shell of the connector mates before the pins make contact on assembly of the connector and breaks after the pins on disassembly. A connector which meets these requirements is available under MIL-C-27599 (8-17) and is the preferred type to be used in rf-proof designs.

The advantages gained using circumferential spring fingers over bayonet coupling is dramatically illustrated in Figure 8-37. In this case, the spring contacts were of silver-plated beryllium copper.

**8.9 SHIELDED ENCLOSURES (SCREEN ROOMS).** Screen rooms are specially constructed enclosures designed to provide an electromagnetically quiet area. In very high level signal environments or where very sensitive equipments must be protected, screen rooms may be necessary. Table 8-19 summarizes some of the more significant features of twelve different types of screen rooms. These same rooms with carefully engineered apertures and openings can be expected to provide at least 100 dB attenuation to electric and plane wave fields. When the installation of a shielded room is required, a number of alternatives must be considered. The most important of these alternatives is whether to shield an existing or future room or building, or whether to provide a demountable enclosure which may be relocated quite simply when the need arises.



Table 8-18

Connector Application Summary

Connector Series	For RG-/U Cables	Disconnect Style	Voltage Rating	Characteristic Impedance	Frequency Range	Method of Assembly
N	5, 6, 8, 9, 10, 11, 12, 13, 14, 17, 18	Screw-on type	500 V peak	50 ohm 70 ohm	Up to 10 GHz	Manual
GR-874	8, 9, 29, 55, 58, 58A, 59, 62, 116	Push-on type	1500 V peak	50 ohm	Up to 7 GHz	Manual
C	8, 9, 10, 12, 14, 55, 58	Bayonet Lock type	1000 V peak	50 ohm	-----	Manual
UHF	8, 9, 10, 11, 12, 13, 44, 58, 62, 63, 65, 71	Screw-on type	500 V	(nonconstant)	Up to 200MHz	Manual
LC	17, 18	Screw-on type	5000 V peak (modified to 10 kV)	50 ohm	-----	Manual
3N	8, 9, 10, 17, 18	Screw-on type	5000 V peak	50 ohm	-----	Manual
3N	55, 58, 59, 62, 71	Screw-on type	250 V peak	(nonconstant)	Up to 200 MHz	Manual
3NC	55, 58, 59, 62, 71	Bayonet lock type	250 V peak	50 ohm	Up to 10 GHz	Manual & Crimp-on
Sub-miniature	174	Screw-on, & Push-on types	-----	50 ohm 75 ohm	-----	Crimp-on

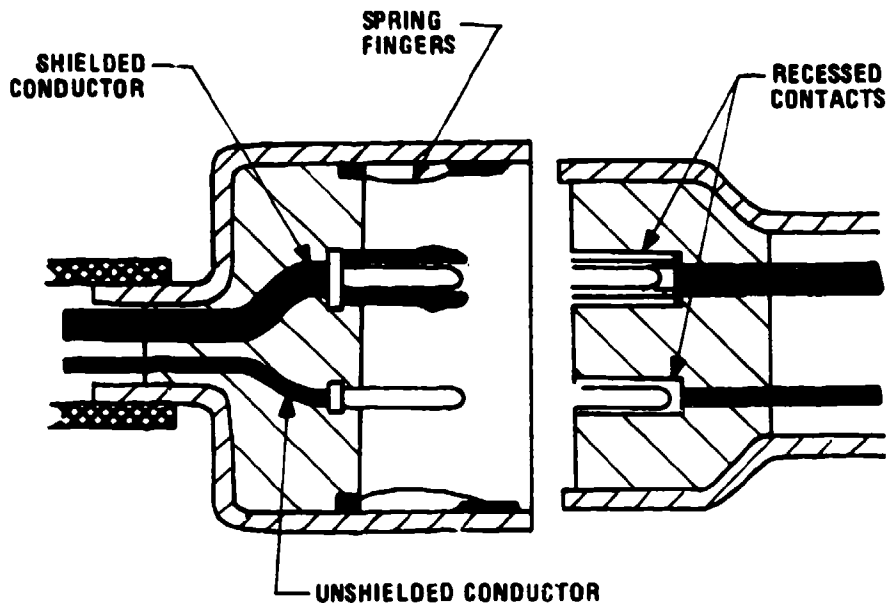


Figure 8-35. Connector for Shield Within a Shield

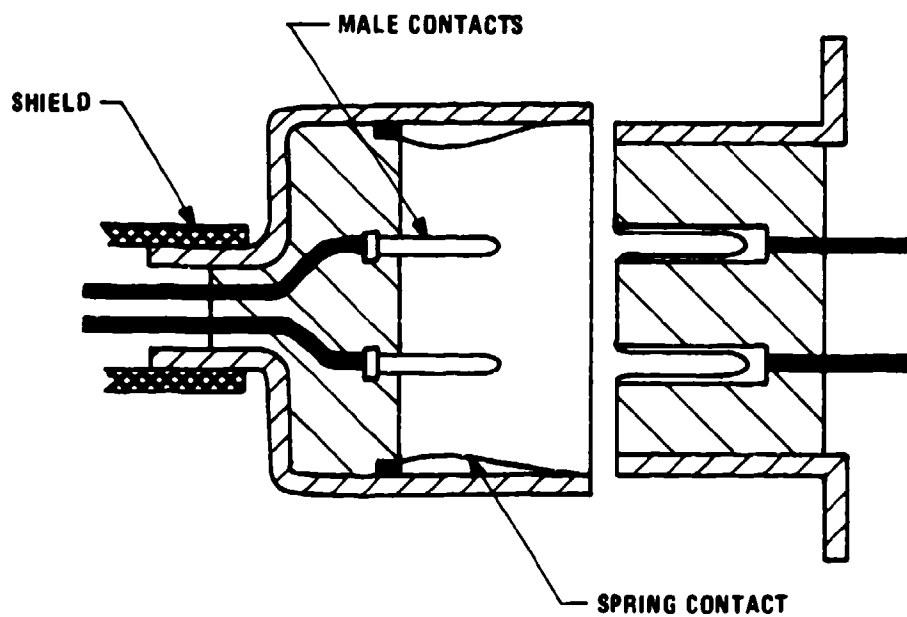


Figure 8-36. RF-Shielded Connector

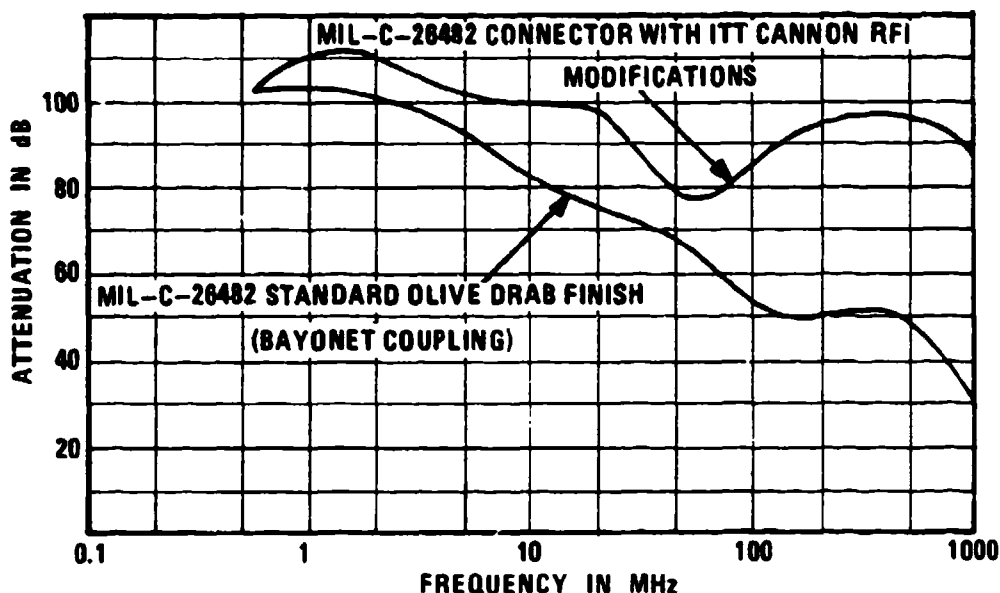


Figure 8-37. Effectiveness of Circumferential Spring Fingers for Improving the Shielding of a Connector (8-18)

#### 8.9.1 Demountable (Modular) Enclosures.

The basic construction of a demountable enclosure might be a 1.27 cm (1/2 inch) thick plywood panel faced on both sides with an electro-galvanized steel sheet of nominal 0.56 mm (0.022") thickness. For non-isolated double shields, the double facing of the walls makes panel-to-panel joining a considerably more certain process as each bonding joint is duplicated. The joining between wall panels is effected by a specially formed metal section, and the design of this requires a fairly precisely controlled blend of resilience and rigidity to establish continuous contact without gaps throughout the length of each bonding member.

The most critical part of any shielded enclosure is the door; with some modern installations doors sizes of 1.86 square meters (20 square feet) and above are required. In general, two types of door bonds are used; these are referred to as the "wedge" and the "knife edge" design. The most commonly used is the wedge door (Figure 8-38) which takes the form of a standard casement type hinged opening leaf or leaves with the frame and the door leaf edges shaped to form a wedge entry, and beryllium copper finger stock affixed in a double layer around the complete periphery of the door leaf. The reason for adopting the wedge design is that, by correctly choosing the angle of wedge, contact pressure on the finger stock can be made high without the risk of tearing and breaking of the spring fingers when the door is opened; it has been found that this type of construction can achieve an overall performance on the order of 125 dB attenuation. The second type of door which has been used for special applications is the knife edge design in which the door leaf is provided with a flanged edge made to enter between two sets of finger stock contacts, enclosed within a channel section fixed upon the door frame. An advantage of this construction is that finger stock is completely protected and the performance is better than obtainable with a wedge door, especially at low frequencies.

Table 8-19

## Characteristics of Commercially Available Shielded Enclosures (8-13)

<u>Room Description</u>	<u>Frequency (kHz)</u>	<u>Magnetic Shielding Effectiveness (dB)</u>	
		<u>MAX</u>	<u>MIN</u>
Copper screen cell type	15	61	56
Styrofoam core, sheet metal skin.	200	97	96
Braided gasket material on door.	15	90	54
	150	87	63
Hollow core construction. Piano	15	100	81
hinge on door with finger stock.	200	118	108
29 mil sheet metal bonded to 3/4" plywood base panel (2 sides) with bolted seam clamps. Three point suspension of personnel door. 20/50 foot overhead door with double row finger stock.	1000	100	80
Construction similar to above,	18	93	64
except no overhead door.	150	120	95
26 gauge steel with folded and soldered seams between panels. Commercially available door with double row of beryllium copper finger stock. All power lines provided with filters.	14	58	34
	280	75	58
Continuously soldered 20 gauge sheet metal with 1.25 oz/ft <sup>2</sup> zinc electroplate. Two commercial doors with finger stock (2 rows). Power line filtering installed. Room size = 20 x 20 x 8 feet.	14	70	65
	100	-	90
Continuously inert-gas welded sheet steel, 12 gauge with overlapping seams. Standard commercial shielded room door with double row finger stock.	14	90	74
	200	130	106

Table 8-19 (Continued)

<u>Room Description</u>	<u>Frequency</u> (kHz)	<u>Magnetic Shielding Effectiveness</u> (dB)	
		<u>MAX</u>	<u>MIN</u>
Similar to above in construction features.	15	111	50
	100	99	81
Double shielding of 10 gauge continuously inert-gas welded low carbon sheet steel, 2" spacing between walls, pneumatic bladder, expanding panel sliding doors (no gasket).	0.1	-	25
	1	62	52
	15	108	92
	100	120	107
Room partitioned into three separate rooms; two are 12' x 12' x 10' and the third is 12' x 12' x 14'. All seams continuously inert-gas welded 16 gauge sheet steel. Doors have pneumatic bladder with triple row of finger stock.	0.5	104	73
	1.0	122	-
	5.0	80	-
	10	39	-
	15	20	-
Room divided into three cells. Single shielding sheet steel continuously inert-gas welded, with pneumatic bladder, and expanding panel sliding doors (EMI gasket for contact surface). Total room size 30 x 70 x 12 feet.	1	115	-
	15	114	104
	100	140	114
	10,000	119	61

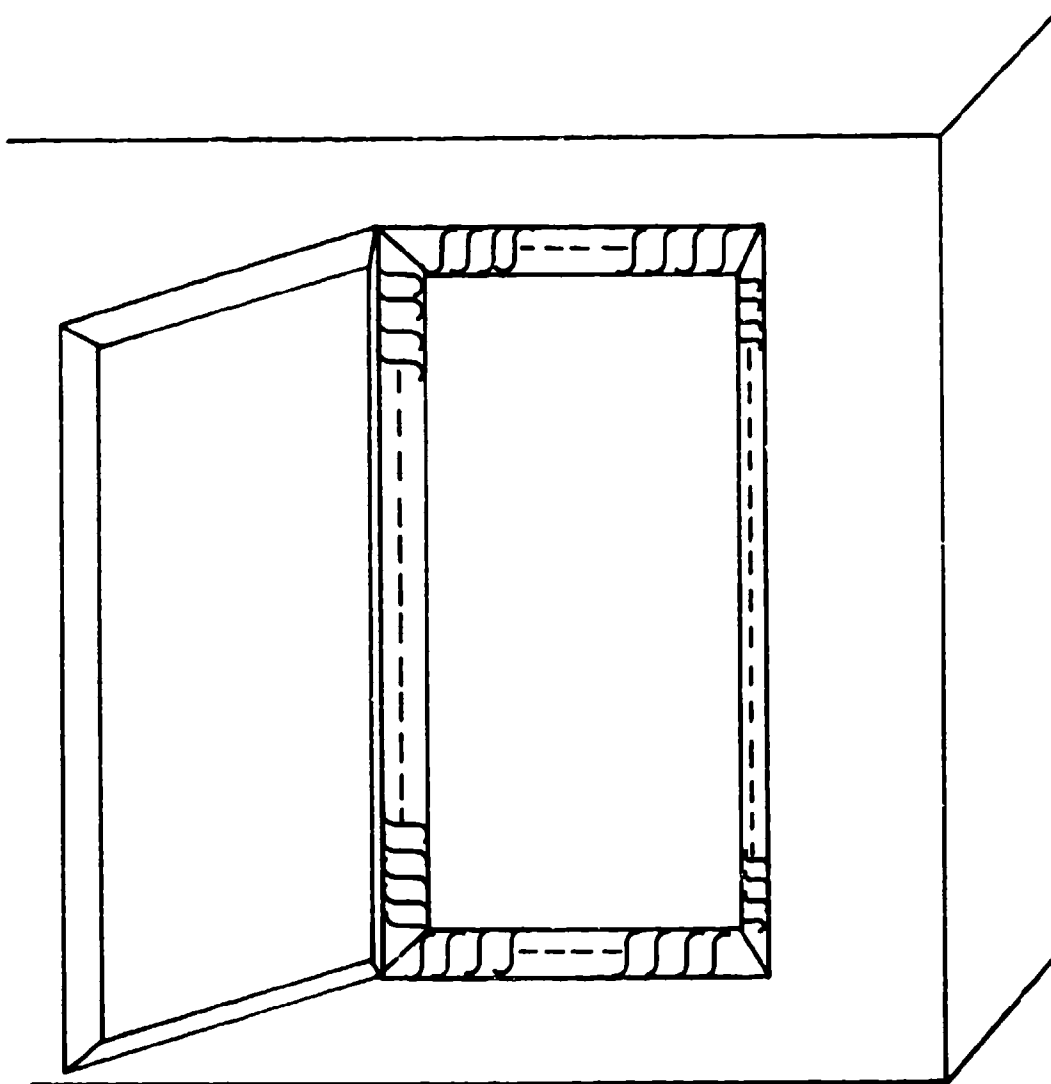


Figure 8-38. Use of Finger Stock for Door Bonding

Additional door bonding may be incorporated with either woven Mumetal gasketing (for very low frequencies), or flexible microwave absorber (for very high frequencies).

To attenuate signals below 50 MHz, waveguide hallways can be used (8-19). The cutoff frequency is proportional to the largest lateral dimension of the hallway; therefore, a tradeoff is generally necessary between hallway size and required attenuation. As shown in Section 8.5.3.1, the amount of attenuation of frequencies below cutoff is a function of hallway length. The waveguide hallway may be constructed of 20 gauge, or thicker, low carbon steel supported by any structurally sound, but nonconductive material.

In all types of door design intended for use at frequencies above a few hundred megahertz, it is desirable to avoid metallic penetration of the door. A special locking catch has been designed which enables full retention of the door leaf and release of the latch from both sides of the door without the need for any metallic penetration of the shield. This lack of metallic penetration is important since even with the most adequate bonding any operating shaft severely increases the risk of shield degradation at frequencies where the shaft's length becomes resonant. It is also important to ensure that even insulating penetrations through the shield which pass through waveguide-below-cutoff tubes are correctly designed. Although the cutoff frequency of a waveguide in air can be easily calculated, the inclusion of insulating material of high dielectric constant in the waveguide considerably reduces the cutoff frequency.

A further requirement for shielded enclosures is adequate ventilation. Honeycomb structures provide a virtually unimpeded passage for air flow and are normally incorporated in ventilation ducts, ventilation openings, and fans or air conditioner systems.

It is essential to avoid signal penetration via power and signal wiring. This demands that filters achieving adequate insertion loss are installed in all incoming cables; it is fairly normal to have three-phase power circuits and several hundred signal lines going into a large enclosure. It is essential that the filters provide the specified attenuation under full-load conditions at all frequencies. Unless the filter attenuation is maintained at all frequencies and load currents, the overall shield attenuation will be degraded by the signal penetration via the filters. Shield penetrations may also be provided for air, gas, and water lines; these can be achieved either by the use of waveguide-below-cutoff tubes carrying insulating piping or by welding metal pipework to the shield. It is essential that all input circuits and penetrations occur in a localized area.

It is necessary that the shield be grounded adequately for safety purposes. Although an external ground connection has no effect on the equipment placed within an ideal shield since the shield itself forms its own private world, an external ground is essential to prevent the enclosure from reaching dangerous potentials relative to its surroundings.

#### 8.9.2 Custom Built Rooms.

In spite of the wide range of use of demountable modular enclosures, a considerable demand exists for specialized custom built shielded areas. These are employed either where the insertion loss requirements are markedly different from those obtainable from modular rooms or where the area to be enclosed is exceptionally large and economy dictates that some other design be adopted. Many forms of construction are used and these include enclosures made from woven copper or steel mesh, from pierced and expanded metal, from aluminum or copper foils, from high permeability materials such as Mumetal, and from all-welded steel sheet.

The use of mesh and open work materials is only employed where a very economical construction is required and only a low shielding performance is necessary. Likewise the high permeability foils are not normally employed, although the low frequency performance of these can be extremely good when related to the foil thickness. A more economical construction often results from the use of welded steel in thicker gauges, although high permeability materials are required where the shield must provide high attenuation to extremely low frequency or constant magnetic fields.

The most efficient practical shielding is provided by a continuously welded steel sheet clad enclosure. Standard practice in Great Britain is to employ a 1.2 mm (0.048") thick electrogalvanized mild steel sheet continuously seam welded along all edges using an inert gas shrouded electric arc welding process. This approach may achieve the highest performance realizable at an economical price. Construction may either be supported by the walls and ceiling of the parent room, or the shield may incorporate its own independent steel framework.

The shielding effectiveness of a shielded enclosure can be improved with the use of double shields. As indicated in the earlier section on the theory of shielding, the shielding effectiveness of two parallel (but slightly separated) shields is better than that of one double thickness shield but not twice as effective as a single-thickness shield. The actual improvement in shielding efficiency is dependent upon the degree of electrical isolation maintained between the two shields.

At least one manufacturer (8-20) of shielded rooms maintains that the isolated double shielded room is substantially more effective than either the single-shielded or the "not isolated double shielded" room. The same types of doors, ventilation apertures, and filters described for the modular rooms are used except that in many cases an rf-proof access lock is provided; this may combine interlocks between the doors and completely automatic operation either by electric, hydraulic, or pneumatic systems.

#### 8.9.3 Foil Room Liners.

When the shielding requirement does not justify an all-welded steel room or a separate screen room, it may be possible to use metal foils. For example, a copper foil nominally 5 mils thick with continuous soft soldered seams may be employed. This copper foil can be glued to the walls, floor, and ceiling to provide a complete lining to an existing room. If this construction is used in conjunction with gasketed metal doors, properly designed vents, and electrical filters, performance, while not being good for low frequency magnetic fields, can be comparable to welded steel at the higher frequencies. To achieve this performance, it is essential that all seams and joints be carefully soldered to establish continuous bonds. The cost of construction is not as low as it might first appear, especially when the additional complications which result from the need to provide fixtures for internal decorative finish and equipment mounting within the shielded area is considered. In general, this form of construction is only used where a relatively unsophisticated enclosure is required, e.g., in certain electro-medical work. If even more economy is required, it is possible to omit the soldering of the joints between the copper foils and use a conductive adhesive tape which is less expensive to install. If only electric fields are present at low frequencies, then a copper foil shield constructed in this manner will probably be adequate.

When shielding is required only for microwave frequencies, a very economic shield may be constructed using aluminum foil of approximately 5 mils thick glued to the walls, floor, and ceiling. An overlap between adjacent foil sheets of approximately 3 cm (2 inches) should be allowed, these overlaps should be secured with aluminum-backed contact adhesive tape. This type of shield is most effective at frequencies above several



hundred megahertz; its shielding effectiveness increases with frequency since the bond between adjacent sheets is primarily capacitive. The normal application for this type of shield is for the protection of computers and data processing installations operating in the vicinity of high power radars. Where shields of this type are intended to work only at very high frequencies, it may be possible to dispense with the shielding over part of the central floor area in ground level installations.

#### 8.10 TESTING OF SHIELDS.

Shield testing may be categorized as (1) the testing of shielding materials to determine their shielding properties, and (2) the testing of shield designs (such as shielded enclosures) to determine whether or not the design and construction are satisfactory. The first category of testing results in design data such as that described earlier in this chapter, and is usually performed by the shielding material manufacturer rather than the equipment designer or user. Methods for performing these and related tests can be found in Reference 8-21 and are not discussed further here. On the other hand, the second category (the testing of equipment shields and shielded rooms for verification of sufficient shielding effectiveness) is a necessary part of equipment development and/or acceptance and is therefore discussed in the following.

The testing of constructed shields is necessitated by the somewhat unpredictable effects of both intentional and unintentional openings and seams in the shield. Localized testing can point out the location of electromagnetic (EM) leaks such as those resulting from welding faults in seams and from poorly fitting gaskets. Such testing is frequently necessary for the successful construction of shielded enclosures. Uniform field (as opposed to local) testing is useful for acceptance testing of a shield. Methods have been developed for both localized and uniform shielding tests for variable-frequency EM fields of low impedance (magnetic), high impedance (electric), and plane waves.

The variety of test methods available for evaluating shielding effectiveness are due, at least in part, to the many different factors that can affect material shielding capabilities. These factors include the configuration of the shield (Is it a sheet of material, or is it a box?), the frequency range of concern, whether or not the impinging wave is planar, the wave impedance, and others. This section will discuss some frequently employed and generally applicable shielding effectiveness tests. Frequently employed tests include:

- a. Low Impedance Magnetic Field Testing Using Small Loops,
- b. Low Impedance Magnetic Field Testing Using a Helmholtz Coil,
- c. High Impedance Electric Field Testing Using Rod Antennas,
- d. High Impedance Electric Field Testing Using a Parallel Line Radiator,
- e. Plane Wave Testing Using Antennas,
- f. Plane Wave Testing Using a Parallel Plate Transmission Line, and
- g. MIL-STD-1377 Testing (8-22).

A number of the above tests are very similar to tests designed to measure equipment and system EMC in accordance with MIL-STD-462 (8-23). They also are similar to tests performed to evaluate EM effectiveness of shielded enclosures used for testing purpose in accordance with MIL-STD-285 (8-24). One who is concerned with the measurement of shielding properties should become familiar with both of these standards.

The MIL-STD-1377 tests represent procedures for evaluating the shielding (and filtering) effectiveness of systems. The specification contains a unique approach to shielding measurements; its cable effectiveness evaluation methods are good illustrations of how cable and connector performance tests should be performed.

It should be pointed out that a high degree of measurement accuracy cannot generally be expected for shielding tests. Typically, wave impedances are not established when the tests are performed, antenna correction factors used for calibration purposes are based on plane-wave assumptions even though the test condition may not warrant this assumption, the degree of radiated field distortion by proximal structures is not known, and other factors limit the accuracy of the measurement. However, the tests can be expected to provide guidance on the shielding design approaches and the general effectiveness to be expected of those approaches.

#### 8.10.1 Low Impedance Magnetic Field Testing Using Small Loops.

This test is designed to indicate the shield's effectiveness in reducing the intensity of predominantly magnetic field radiation. It employs two small loop antennas and evaluates loop coupling with and without an intervening shield. MIL-STD-285 incorporates a similar magnetic field small loop measurement procedure to evaluate the shielding effectiveness of shielded enclosures used for electronic testing purposes.

In this test, a pair of identical small loop antennas are used, one on one side of the shield and one on the other, spaced equidistant from the shield. If an enclosure is being tested, the usual practice is to have the test signal source within the enclosure and the receiving loop and detector outside the enclosure.

Figures 8-39 and 8-40 show the two basic loop orientations. In Figure 8-39 the loops are coaxial, that is, both loops are normal to a common loop axis. In Figure 8-40 the loops are coplanar, that is, the loop surfaces lie on the same plane. Tests using at least these two orientations should be employed, but orientations that may result in a lower effectiveness figure should not be ignored. Both the loop diameters and the loop separations should be significantly less than the shortest dimension of the box, container, or enclosure being tested. Since this will result in only a small section of the shield being illuminated at one time, it will be necessary to move the loop over the entire surface of the shield to establish the effectiveness of the shield.

The frequency range over which this test can be performed is a function of the level of shielding effectiveness that must be measured (measurement system dynamic range), the sensitivity of the test equipment, the available power to drive the test transmitting loop, and the loop-to-shield separations. The limiting factors are usually the areas of the loops and the number of turns in the loops, since these establish the self-resonance frequency of the loop. Loop-to-loop separation should not be closer than the loop diameter.

The small loop-to-loop setup specified in MIL-STD-285 is shown in Figure 8-40 with the following parameter values employed:

Loop diameters (d): 12 inches (30.48 cm)

Loop-to-shield separations (r/2): 12 inches

Loops: One turn of No. 6 AWG Copper Wire

The test setup in this specification is intended to provide a minimum of 70 dB measurement range.

The corresponding test for a uniform magnetic field requires placing the test shielded enclosure within a Helmholtz coil (large loop), with a small detection loop inside the test enclosure. The use of a Helmholtz coil enables a large portion of the enclosure to be illuminated at one time. Various orientations of the sample relative to the loop should be tried. The frequency range is limited by the test sample size, which affects the size of the Helmholtz coil. Increasing coil size increases its inductance, reduces its self-resonant frequency, and decreases the frequency range over which its magnetic field strength remains constant. The coil diameter should be at least two and preferably three times the longest test sample dimension. The upper frequency limit is typically 100-500 kHz.

8.10.2 Additional Test Methods. Although additional test methods for medium and high impedance incident waves exist, they are less frequently applied since the most difficult problem is the shielding at low frequency, magnetic fields.

#### 8.11 PERSONNEL PROTECTION SHIELDS.

Shields for the reduction of EMI are also appropriate for protecting people from potentially hazardous radiation, either ionized or nonionized in nature. For most situations where sensitive electronic apparatus is present, the facility shielding required to prevent EMI is more than adequate to provide personnel protection. However, for high level incident fields, the need for personnel protection alone should not be overlooked. Areas of particular concern are those near high voltage vacuum devices which may emit X-rays, near high power rf sources or emitters such as acquisition and search radars, or near other sources of potentially damaging emanations such as laser emissions encountered during maintenance of fiber optics containing laser diodes.

Shields for protection against contact with hazardous voltages at very low frequencies, i.e., dc and 50/60 Hz, are not generally of the same type as those which protect against radiated fields. Personnel protection may be nonconducting and function more as a simple physical barrier which prevents accidental contact. Metal electromagnetic shields may also establish a physical barrier; however, the barrier is an incidental byproduct and should not be considered to be the primary purpose of the shield.

8.12 DETERMINATION OF SHIELDING REQUIREMENTS. Comprehensive shielding of a structure, particularly a large one, can be very expensive. Fortunately, if the threat signal environment is known or can be predicted, an appropriate choice of available or existing materials can accomplish the necessary shielding with minimum costs. Methods available for establishing the amount of shielding required in a given location include analyzing equipment malfunctions or disturbances, performing an electromagnetic site survey to obtain power density levels, and performing electromagnetic susceptibility and emissions tests of the equipments which are to be located in the facility. Shielding requirements can then be determined by comparing the susceptibility levels of the equipment against the power density levels measured in the area where the equipment is to be located.

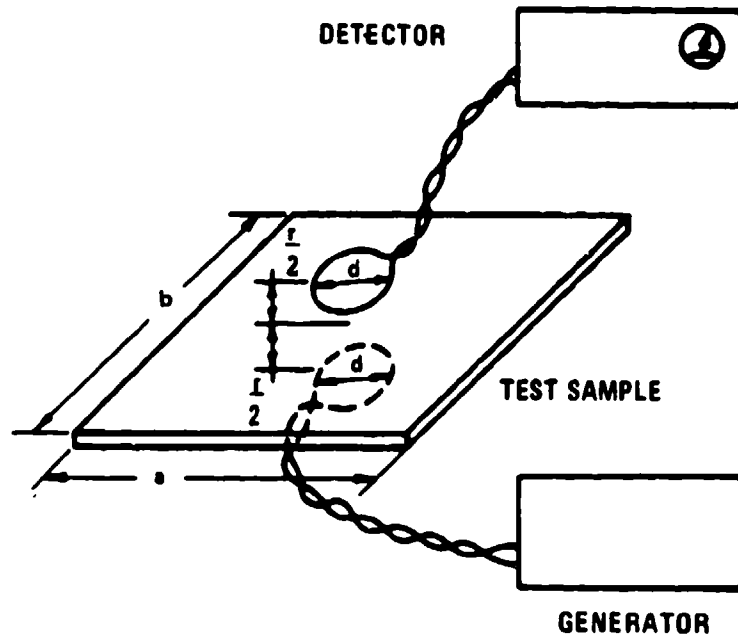


Figure 8-39. Coaxial Loop Arrangement for Measuring Shield Effectiveness

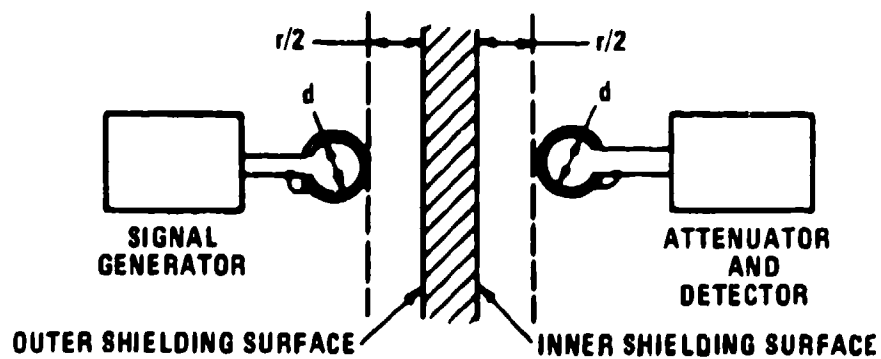


Figure 8-40. Coplanar Loop Arrangement for Measuring Shield Effectiveness

### 8.12.1 Equipment Disturbances.

A reliable indicator of the need for shielding of an equipment is the degree of interference that it experiences or causes. Recognizing that interference can be the result of one of the four different coupling modes, it must be determined that coupling will occur through one of the modes which can effectively be combatted by shielding. For example if the interfering signal is coupled into the equipment or system on a power or signal line, shielding the equipment may accomplish little. The line picking up the disturbing signal may be made less susceptible to interfering signals by careful shielding of the line itself. If inductive, capacitive, or radiated coupling is the cause of the problem, then shielding of the cable either alone or along with the equipment will be effective.

If the equipment is going into a new facility and the decision to be made is whether or not shielding is necessary, the behavior of that equipment in other similar environments should be considered. If the performance of the specific equipment is not known, the behavior of equipments of similar types or construction should be studied. The most reliable method of determining shielding requirements is to compare known susceptibility levels of the equipment or system with known measured power density levels in the area where the equipment or system is being installed.

### 8.12.2 Electromagnetic Environmental Survey.

The most effective way of determining the power densities at the location where the equipment or the structure is to be located is by conducting an electromagnetic environmental survey. This survey is performed using calibrated antennas with special field strength meters or spectrum analyzers. These instruments permit the strength of radiated fields to be determined in terms of volts per meter or in power density, i.e., watts per square centimeter or square meter. For personnel hazard determination, commercially available rf radiation monitors may be used.

The spectrum survey should attempt to identify the presence of all potentially interfering fields. Of particular concern is the field strength of the signals emitted by readily identifiable sources such as commercial radio and television stations, and radar and communications transmitters. Other possible sources of interference include rf heating units, rf welders, microwave ovens, and, in locations near medical facilities, diathermy and electrocautery machines. Desk top evaluations can also be employed to calculate power density/signal strength levels in a given area if all local emitters (including output power, locations, etc) can be identified.

The electrical power system can also be a source of interference. High voltage transmission systems, in particular, frequently generate noise through corona discharge and arcing across dirty connectors and insulators. The frequency spectrum of this noise generally extends well into the HF region (3-30 MHz) or above and can be a cause of severe problems. The routing, either existing or planned, of power lines should be noted carefully. If long runs of signal and control cables in parallel with power lines, either overhead or underground, are unavoidable, shielding of the signal and control cables may be necessary.

In addition to the above identifiable sources of energy against which shielding may be required, other less obvious sources exist. For example, ignition noise from internal combustion engines can be troublesome. Also, office machines, vending machines, and fluorescent lights have been frequently observed to produce interference in digital computers, measuring systems and other sensitive equipments.

### 8.12.3 Equipment EMI Properties.

Different equipments will exhibit different emission and susceptibility properties depending upon the job to be performed, the method of design, the type construction, the type components used, and a variety of other factors. The best indicator as to how much shielding is going to be required for a given piece of equipment or for an entire complex is provided by the measured level of emissions or the susceptibility level of the equipment or system. These properties are determined by operating the equipment in an electromagnetically controlled environment and by (1) measuring the frequency and amplitude of the signals radiated or produced by the equipment or (2) irradiating or otherwise subjecting the equipment to a known field or given signal and noting the minimum level to which the equipment or system responds. Under field conditions, neither of these procedures should be expected to provide precise detailed data because reradiation and mutual coupling effects can cause wide variations in the measured results. However, with a reasonable sampling of the fields or with illuminations provided at various locations and different orientations, an order-of-magnitude estimate of the relative susceptibility or threat posed by the equipment or system should be possible. If precise data is needed, test procedures in accordance with accepted standards, such as MIL-STD-461 and MIL-STD-462 should be performed. Unfortunately because of the expense of performing detailed and accurate emission and susceptibility tests of equipments (even the ability to perform these tests on large complexes in a meaningful manner is doubtful), and because a decision is frequently required on structural shielding before the specific equipment population is known, it is generally necessary to direct attention only to the most critical equipments or systems expected to be installed in the facility. Shielding requirements can also be determined by comparing the susceptibility levels (MIL-STD-461) of the equipment being installed with the measured or calculated power density levels in the area where the equipment is being installed.

If it is simply not possible to anticipate or project the shielding requirements, then the resultant electromagnetic environment in which equipment will be required to perform must be measured or calculated and the information provided to the equipment supplier so that appropriate steps can be taken to assure that the equipment or system will function in that environment.

### 8.13 SYSTEM DESIGN CONSIDERATIONS.

The total area or volume of a facility to be shielded and the physical configuration of the shield is a function of:

- a. the size of the equipment or system requiring shields;
- b. the physical layout including orientation between sources and receptors;
- c. the amplitude and frequency of the interfering signals; and
- d. the cost of materials.

These factors typically interact and, although in a given situation one will predominate, all must be considered.

**8.13.1 Size.** If a very sensitive piece of equipment or small system is to be located in a large structure, shielding the entire structure to protect that one small element is probably not cost effective. The cost of shielding is closely related to the size of the enclosed volume, assuming all other factors equal. Thus, a more economical approach would perhaps be to shield only the room in which the equipment is to be located, construct a shielded cage just for the susceptible (or offending) equipment, or upgrade the shielding of the particular equipment cabinet or enclosure. If, on the other hand, the susceptible element is a fairly large system, e.g., a communications center or a large scale computer, then incorporating appropriate shielding materials into the walls, floor, and ceiling of the room or structure may be necessary. If this requirement is recognized early in the design stage of the facility, the required shielding may be provided by properly-installed conventional structural materials. Also, supplemental shields can frequently be installed with greater economy if done during construction rather than later.

**8.13.2 Layout.**

If a susceptible equipment or system is to be located in a building and some choice exists as to position, special effort should be made to take advantage of the inherent shielding properties of the structure. The existence of metal walls, decorative screens, and other conductive objects may provide all the shielding necessary. Further, equipments frequently are more sensitive to radiated signals impinging from only one or two directions. Thus, orienting the equipment such that the susceptible side is facing away from the incident signal can lessen the shielding requirements.

Signal and control cables deserve special mention. Because the voltage (or current) in the receptor wire is inversely dependent upon the distance from the source wire and directly proportional to the length of the path, every effort should be made to avoid long runs in parallel.

**8.13.3 Signal Properties.**

The shielding effectiveness of practically all materials is frequency dependent. The type of shield which will protect against an X-band radar signal will not necessarily be effective against a commercial broadcast transmitter. In choosing a shield for a particular purpose, compare the attenuation properties of the material with the frequency of the threat signal.

The amplitude of the signal to be shielded indicates the amount of field attenuation the shield must provide. For most fields, the attenuation provided by the shield is not influenced by the magnitude of the field, i.e., a shield which will attenuate a low level field 60 dB will likewise attenuate a high level field 60 dB. Very strong magnetic fields, however, can cause saturation effects and the attenuation of the shield will generally decrease under very strong magnetic fields. This phenomenon is very important in choosing shields to protect against EMP for instance. Where saturation effects are likely, thicker shields are required in order to maintain the attenuation needed to protect against the very strong fields.

**8.13.4 Cost.**

The impact of size on cost was noted previously in Section 8.13.1 above. Other cost factors to consider include those associated with providing input and output ports for wiring and cabling, ventilation, and physical and visual access (doors, windows, meter openings, etc.) while maintaining the effectiveness of the shield.

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## CHAPTER 9

## PERSONNEL PROTECTION

9.1 ELECTRIC SHOCK.

Electric shock occurs when the human body becomes a part of an electric circuit. It most commonly occurs when personnel come in contact with energized devices or circuits while touching a grounded object or while standing on a damp floor. The major hazard of electric shock is death. Fatalities from shock total about 1,000 annually. In addition, numerous injuries occur each year due to involuntary movements caused by reaction currents.

The effects of an electric current on the body are principally determined by the magnitude of the current and the duration of the shock. The current is given by Ohm's Law, which, stated mathematically, is  $I=V/R$  where  $V$  is the open circuit voltage of the source and  $R$  is the resistance of the total path including the internal source resistance, and not just the body alone. In power circuits, the internal source resistance is usually negligible in comparison with that of the body. In such cases, the voltage level,  $V$ , is the important factor in determining if a shock hazard exists.

At the commercial frequencies of 50-60 Hz and at voltages of 120-240 volts, the contact resistance of the body primarily determines the current through the body. This resistance may decrease by as much as a factor of 100 between a completely dry condition and a wet condition. Thus, perspiration on the skin has a great effect on its contact resistance.\* At voltages higher than 240 volts, the contact resistance of the skin becomes less important. At the higher voltages, the skin is frequently punctured, often leaving a deep localized burn. In this case, the internal resistance of the body primarily determines the current flow.

9.1.1 Levels of Electric Shock (9-1) (9-2).

The perception current is that current which can just be detected by an individual. At power frequencies, the perception current usually lies between 0 and 1 milliams for men and women, the exact value depending on the individual. Above 300 Hz, the perception current increases, reaching approximately 100 milliams at 70 kHz. Above 100-200 kHz, the sensation of shock changes from tingling to heat. It is believed that heat or burns are the only effects of shock above these frequencies.

The reaction current is the smallest current that might cause an unexpected involuntary reaction and produce an accident as a secondary effect. The reaction current is 1-4 milliams. The American National Standards Institute (9-3) limits the maximum allowable leakage current to 0.2 milliams for portable two-wire devices and 0.75 milliams for heavy movable cord-connected equipment in order to prevent involuntary shock reactions.

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\*For calculation purposes, the resistance of the skin is usually taken to be somewhere between 500 and 1500 ohms.

Shock currents greater than the reaction current produce an increasingly severe muscular reaction. Above a certain level, the shock victim becomes unable to release the conductor. The maximum current at which a person can still release a conductor by using the muscles directly stimulated by that current is called the "let-go" current. The "let-go" current varies between 4-21 milliamps, depending on the individual. A normal person can withstand repeated exposure to his "let-go" current with no serious after effects when the duration of each shock lasts only for the time required for him to release the conductor.

Shock currents above about 18 milliamps can cause the muscles of the chest to contract and breathing to stop. If the current is interrupted quickly enough, breathing will resume. However, if the current persists, the victim will lose consciousness and death may follow. Artificial respiration is frequently successful in reviving electric shock victims.

Above a certain level, electric shock currents can cause an effect on the heart called ventricular fibrillation. For all practical purposes, this condition means a stoppage of the heart action and blood circulation. Experiments on animals have shown that the fibrillating current is approximately proportional to the average body weight and that it increases with frequency.

In Table 9-1, the various hazardous current levels for ac and dc are summarized along with some of the physical effects of each.

Table 9-1

## Summary of the Effects of Shock (9-1) (9-2)

<u>Alternating Current (60 Hz)</u>	<u>Direct Current</u>	<u>Effects</u>
(mA)	(mA)	
0-1	0-4	Perception
1-4	4-15	Surprise (Reaction Current)
4-21	15-80	Reflex Action (Let-Go Current)
21-40	80-160	Muscular Inhibition
40-100	160-300	Respiratory Block
Over 100	Over 300	Usually Fatal

### 9.1.2 Shock Prevention.

Most shock hazards can be divided into two categories: unsafe equipment and unsafe acts. The most common hazards in each category can be controlled as follows:

- a. Power cords and drop cords with worn and/or broken insulation should be routinely replaced.
- b. All spliced cords should be removed from service.
- c. Exposed conductors and terminal strips at the rear of switchboards and equipment racks should be enclosed and warning labels installed.
- d. Rubber mats should be installed on the floor of all enclosures containing exposed conductors and on the floor in front of high voltage switches.
- e. High voltage switches should be of the enclosed safety type.
- f. All wiring should comply with recognized electrical codes and it should be large enough for the current being carried.
- g. Temporary wiring should be removed as soon as it has served its purpose.
- h. The noncurrent-carrying metal parts of equipment and power tools should be grounded.
- i. The main power switch to all circuits being worked on should be locked open and tagged.
- j. Power switches should be opened before replacing fuses and fuse pullers should be used.
- k. Fuse boxes should be locked to prevent bridging or replacing with a heavier fuse.
- l. Care should be taken to prevent overloading of circuits.

### 9.2 STATIC ELECTRICITY.

Static electricity is produced when two bodies, particularly of unlike materials, are brought into intimate contact and then separated. When in contact, there is a redistribution of charge across the area of contact and an attractive force is established. When the bodies are separated, work is done in opposing these attractive forces. This work is stored in the electrostatic field which is set up between the two surfaces when they are separated. If no conducting path is available to allow the charges to bleed off the bodies, the voltage between the surfaces can easily reach several thousand volts as they are separated.

Static electricity is an annoyance to many individuals. Static shock can result in discomfort and even injury to workers due to involuntary reaction. A far more dangerous aspect of static electricity is the fire and explosion hazard. This hazard can occur in situations where a vapor-air, gas-air, dust-air, or combinations of these

mixtures exist in the proper ratio. In order for static to be a source of ignition of the mixtures, four conditions must be fulfilled. These conditions are:

- a. There must first be an effective means of static generation.
- b. There must be a means of accumulating the separate charges and maintaining a suitable difference of electrical potential.
- c. There must be a spark discharge of adequate energy.
- d. The spark must occur in an ignitable mixture.

The most common sources of static electricity are:

- a. Steam, air, or gas flowing from any opening in a pipe or hose, particularly when the stream is wet or when the air or gas stream contains particulate matter.
- b. Pulverized materials passing through chutes or pneumatic conveyors.
- c. Nonconductive power or conveyor belts in motion.
- d. Moving vehicles.
- e. All motion which involves changes in relative position of contacting surfaces (usually of dissimilar substances, either liquid or solid), of which one or both must be a poor conductor of electricity.

Static electricity can be controlled in a variety of ways. The most effective means are:

- a. Bond all metallic parts of a system to prevent the existence of a statically-induced potential difference between any two metallic objects in the system.
- b. Ground all metallic systems to prevent the accumulation of static charges.
- c. Increase the relative humidity to 60% - 70% to increase the moisture content and thus the conductivity of insulating materials such as fabric, wood, paper, concrete, or masonry.
- d. Use ionizing devices to ionize the surrounding air so that it becomes sufficiently conductive to bleed off static charges.
- e. Use conductive materials for rugs, flooring, belts, etc. where nonconductive materials might otherwise be used.

### 9.3 RADIO FREQUENCY (RF) RADIATION HAZARDS.

The effect of rf radiation on living tissue is thought to be primarily thermal in nature. The most vulnerable parts of the human body are the eyes and the testes. However, other parts which can be affected are the brain, nerves, skin, and muscles. The thermal effects can range from a mild heating of the skin or organs to fatal damage.

Below 1000 MHz, rf energy penetrates deeply into the body. These frequencies are extremely hazardous since the radiation is not detected by the nerve endings located in the skin. The power absorbed in the body tissues can be as high as 40% of the incident power. The urinary bladder, gall bladder, and parts of the gastrointestinal tract are particularly vulnerable since they are not cooled by an abundant flow of blood. Also, stainless steel and platinum bone implants and fillings in teeth can increase in temperature when subjected to rf radiation, resulting in burning of tissues.

In the 2-5 GHz region of the rf spectrum, the eyes and the testes are the most vulnerable organs to rf radiation damage. Damage to the eyes is generally irreversible and can result in blindness from cataracts or loss of lens transparency. Animal experiments have shown that damage to the testes from low levels of exposure does not differ from that caused by common forms of heat applied to the testes, and that the reduction in testicular function due to heating appears to be temporary. It is not known if rf radiation produces any genetic damage.

To minimize possible hazards from rf radiation, Dept of Defense Instruction 6055.11 (9-4) provides recommendations to prevent possible harmful effects in human beings exposed to radio frequency radiation.

### 9.4 LASER HAZARDS (9-5).

Biological damage from laser radiation is caused by photochemical, thermal, and pressure effects, acoustic and ultraviolet shock waves, plasma generation, ultrasonic emission, and even the generation of free radicals. Of these, the first three are the most hazardous to tissues, organs, and eyes. The damages include tissue ionization, molecular rearrangements, blood vessel occlusion, corneal opacity, retinal lesion, blindness, and even death. Lasers are divided into four classes: Class 1 is non-hazardous; Class 2 depends on blink reflex for the person to turn away to prevent a hazard; Class 3 is a direct or specular reflection hazard; Class 4 is all other high energy lasers. See ANSI Z136.1 for further information concerning safe use of lasers.

It is believed that damage to eye tissue by visible and infrared light is mainly thermal in nature. The lens of the eye is practically transparent in these regions, thus increasing the susceptibility to retinal burn or lesion. The susceptibility is enhanced by the fact that the power density of light converging on the retina is concentrated by a factor of  $10^5$  when it passes through the pupil and lens of the eye.

At near ultraviolet, the eye responds in nearly the same way as it does to visible light, the exception being a marked decrease in vision between 380 nm ( $10^{-9}$  meters) and 420 nm. This decrease is caused by the strong absorption at the lens of the eye. Extreme exposure to these wavelengths may lead to the development of cataracts. In the B and C ultraviolet bands, radiation is absorbed by the cornea and its outer layer. Excessive exposure to these wavelengths can cause a condition called "welder's flash," an effect similar to snow blindness.

Serious skin injury can occur at high power levels in the near infrared and visible regions. The skin becomes increasingly sensitive in the ultraviolet region. Energy at these wavelengths penetrates deeply and can cause severe burns. In the near infrared range, the skin becomes relatively transparent, making the internal organs particularly susceptible.

#### 9.5 X-RAY RADIATION.

X-rays are generated when electrons are accelerated to a sufficiently high velocity before colliding with an appropriate target. In addition to being produced by specifically designed equipment, X-rays can also be produced by high-voltage (> 15 kV) tubes used for other applications. It is important that all sources of X-ray radiation in equipments be identified and shielded so that they will not present a personnel hazard.

The maximum safe exposure to X-ray radiation is considered to be 100 milli-Roentgens per week (9-6). Based on a 40-hour work week, this limit translates into a maximum hourly rate of exposure of 2.5 milli-Roentgens per hour.

#### 9.6 REFERENCES.

- 9-1. C.F. Dalziel, "Electric Shock Hazard," IEEE Spectrum, Vol 9, No. 2, February 1972, pp 41-50.
- 9-2. "Standard General Requirements for Electronic Equipment," MIL-STD-454J, 30 June 1985.
- 9-3. "Leakage Current for Appliances," C101.1-1973, American National Standards Institute, 1430 Broadway, New York NY 10018.
- 9-4. Dept of Defense Instruction 6055.11, 20 August 1986, "Protection of DOD Personnel from Exposure to Radio Frequency Radiation."
- 9-5. Marce Eleccion, "Laser Hazards," IEEE Spectrum, Vol 10, No.8, August 1973, pp 32-38.
- 9-6. Dept of Defense Instruction 6055.8, 3 January 1983, "Occupational Radiation Protection Program."

## CHAPTER 10

## NUCLEAR EMP EFFECTS

**10.1 INTRODUCTION.** In addition to the blast, thermal effects, and radioactive fallout, a nuclear detonation produces an intense electromagnetic effect. Under the proper circumstances, a nuclear detonation generates a high-intensity electromagnetic pulse (EMP) whose frequency spectrum may extend from below 1 Hz to above 300 MHz. This high-intensity EMP can disrupt or damage critical electronic facilities over an area as large as the continental United States, unless protective measures are taken in the facilities. The development of such protective measures involves grounding, bonding, and shielding and requires an understanding of the EMP itself.

**10.2 EMP GENERATION.**

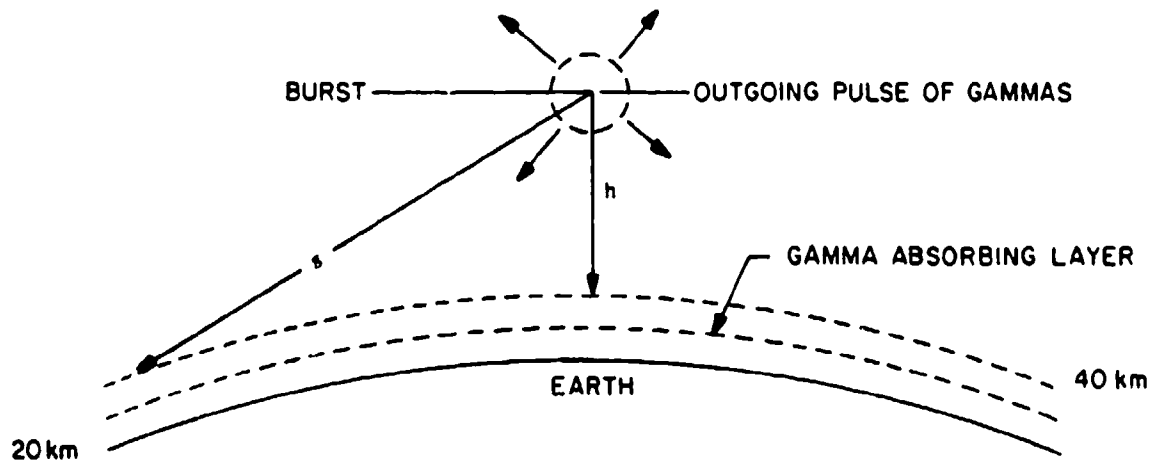
**10.2.1 High-Altitude EMP (HEMP).** The high-altitude EMP (HEMP) produced by an exoatmospheric nuclear explosion is the form of EMP commonly of most interest because of the large area covered by a single bomb. The HEMP is also the form for which interaction and protection are most advanced. The standard HEMP waveforms to be used for tests and analyses of hardened systems are given in DoD-STD-2189 (SECRET-RD). A brief description of the three parts of the standard waveform is given below.

**10.2.1.1 Early-Time HEMP.**

The detonation of a nuclear weapon produces high-energy gamma radiation that travels radially away from the burst center. When the detonation occurs at high altitudes where the mean free path of the gamma photons is large, these photons travel great distances before they interact with another particle. As illustrated in Figure 10-1, gamma rays directed toward the earth encounter dense atmosphere where they interact with air molecules to produce Compton recoil electrons and positive ions. The Compton recoil electrons also travel radially away from the burst center initially, but these moving charged particles are acted upon by the Earth's magnetic field, which causes them to turn about the magnetic field lines (10-1).

The Earth's magnetic field accelerating the Compton recoil electrons causes them to radiate an electrodynamic field. Thus, the early-time HEMP is produced by this charge acceleration (electron turning) phenomenon that occurs in the atmosphere in a region about 20 km thick and 30 km above the Earth's surface (sea level). This source region covers the Earth within the solid angle subtended by rays from the burst point that are tangent to the surface of the Earth, as illustrated in Figure 10-2. To an observer on the ground, the incoming wave appears to be a plane wave propagating toward him from the burst point. The amplitude, duration, and polarization of the wave depend on the positions of the burst and the observer, relative to the Earth's magnetic field lines. Peak electric field strengths of over 50 kV/m with risetimes of a few nanoseconds and decay times of less than 1  $\mu$ s are typical for this early-time portion of the HEMP (10-2).





$h$  = HEIGHT OF BURST = 400 km = 250 MILES

$s$  = DISTANCE TO HORIZON = 2,250 km = 1,400 MILES

A HIGH ALTITUDE BURST ILLUMINATES LARGE GEOGRAPHICAL REGIONS WITH GAMMA RAYS.

Figure 10-1. EMP from High Altitude Bursts

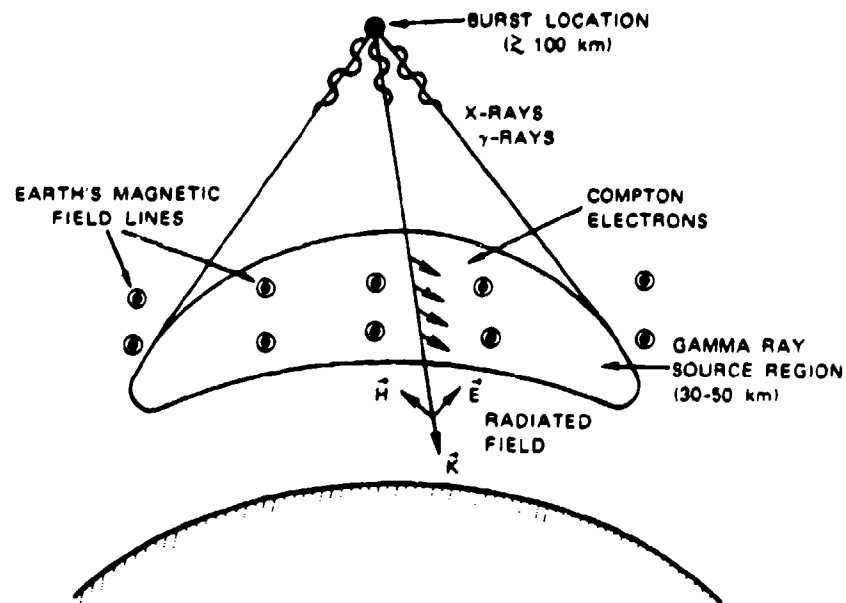


Figure 10-2. Schematic Representation of High-Altitude EMP Generation

**10.2.1.2 Late-Time HEMP (MHDEMP).** Much later (0.1 to 100 s), currents are induced in the ground by the effects of the expanding and rising fireball constituents. These effects are called the magnetohydrodynamic EMP (MHDEMP). They arise from the motions of the rapidly expanding bomb debris and hot ionized gases in the Earth's magnetic field. MHDEMP has two phases produced by two principal effects. The first effect is an ionospheric blast wave that deforms the geomagnetic field lines and produces an early phase of the MHDEMP that reaches the Earth's surface in 2 to 10 seconds and can be seen worldwide. The second effect is the "atmospheric heave," in which hot debris and air ions are moved across geomagnetic field lines to cause large circulating currents in the ionosphere. These currents induce image currents in the ground over a period of 10 to 100 seconds. Although the field strengths produced at the surface by the MHDEMP are small (tens of volts per kilometer), they occur over long times. Thus, the MHDEMP is a consideration for long power and communications lines and, because of its duration, for the energy it can deliver to protective devices.

**10.2.1.3 Intermediate-Time HEMP.** Between the early-time HEMP and the MHDEMP, transitory phenomena produce what is called intermediate-time HEMP. This HEMP lasts from about 1  $\mu$ s to about 0.1 s. The intermediate-time HEMP observed at the Earth's surface has a peak electric field strength of a few hundred volts per meter and is predominantly vertically polarized.

**10.2.2 Surface-Burst EMP.** When a nuclear weapon is detonated at or near the surface of the Earth, neutrons and gamma rays are ejected radially outward from the burst center. The gamma ray photons emitted by the bomb, and others produced by neutron inelastic collisions with air, ground, and water, interact with air molecules to produce Compton recoil electrons. At or near sea level, however, the Compton recoil electrons quickly collide with air molecules to provide a copious supply of low-energy secondary electrons and ions. Thus, the Compton recoil electrons account for a large charge separation and, because of the secondary ionization, a fairly conductive air. As illustrated in Figure 10-3, the charge displacement is asymmetrical because of the Earth's surface. The initial dipole charge is discharged by current through the ionized air and soil. From a large distance, the EMP from a surface burst appears to emanate from a dipole source; it is vertically polarized and attenuated as  $1/r$  with distance,  $r$ , from the burst point. Thus, the surface-burst EMP is a more localized source than the HEMP. However, within the source region where the Compton electrons, secondary ionization, and relaxation currents occur, the fields are large, and long conductors, such as power lines and communication cables, may have large currents induced on them. These currents may be propagated along the conductors for great distances from their source. Therefore, this source-region EMP (SREMP) may be important to systems far outside the source region if they are connected to the source region through wires, cables, or other conductors.

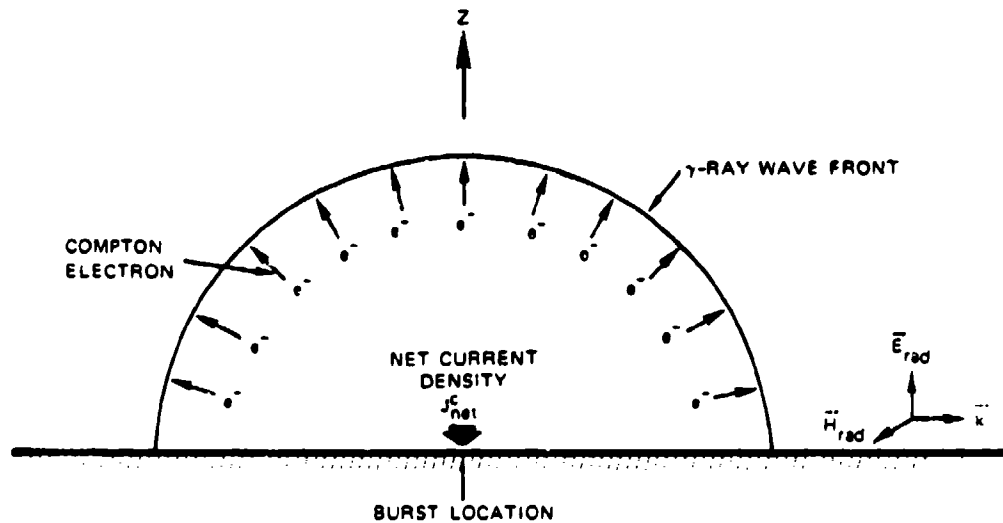


Figure 10-3. Surface-Burst Geometry Showing Compton Electrons and Net Current Density,  $J_{net}^C$ . Radiated Fields are Approximately Proportional to  $dJ_{net}^C/dt$  (Electric-Dipole Fields)

### 10.2.3 Other EMP Phenomena.

The high-altitude EMP (HEMP) is, by far, the most important form of EMP for communication facilities because of its large area of coverage. However, in addition to the HEMP and the surface-burst EMP, a few other electrical effects should be mentioned. System-generated EMP (SGEMP) is produced when the high-energy particles (mostly gamma- and X-ray photons) produced by the bomb interact directly with the system structure. These interactions knock electrons out of the structure, which causes current on the structure and potential gradients between the structure and the removed charge. The structure of interest may be system wiring or cable shielding; the current and potential differences are then on system circuits. (Because this EMP is often generated inside the system, it is sometimes called internal EMP (IEMP).) SGEMP is of major concern for satellites and other space vehicles because the gamma- and X-rays from the high-altitude bomb can travel great distances without colliding with another particle or structure. SGEMP is also a consideration for surface systems whose blast and thermal resistance permits them to operate inside the source region.

Another important electrical effect is known as transient radiation effects on electronics (TREE). The radiation emitted by the nuclear explosion can interact with components of electronic circuits to produce ionization or atomic displacements in the semiconductor and insulating materials. The effects range from momentary changes in conductivity to permanent changes in crystal lattices. Semipermanent effects, such as trapped charges in insulating materials, may also occur. TREE may upset memories, produce spurious circuit responses (logic errors), drive circuits into abnormal states, or cause permanent damage. As with most other EMP forms, damage caused by TREE can also occur through secondary effects. Self-inflicted damage may be triggered by abnormal conductivity in a junction that allows stored energy to be released. In addition, one circuit may be caused to instruct another circuit or another part of the system to perform some forbidden act that destroys the circuit or even the system.

#### 10.2.4 Comparison with Lightning.

Lightning and the EMP are often compared because they are both large electromagnetic phenomena and because more people have experienced lightning in some form. Though they are generated by different mechanisms, some aspects of their effects on systems are similar. Both can produce large electrical transients in systems. Both interact with power lines and communication cables to excite systems served by these cables.

However, lightning and HEMP have important differences in their electromagnetic properties and in the way they interact with systems. Lightning can deliver greater energy to a moderate impedance load, such as a power transmission line, than can the HEMP. On the other hand, the HEMP has a larger rate of change of field and induced currents and voltages than lightning, so that coupling phenomena that depend on  $dE/dt$  and  $dB/dt$  (where E and B are the electric field intensity and magnetic flux density, respectively) are more important for the HEMP excitation than they are for lightning. Because the HEMP appears to be a plane wave at the Earth's surface, its interaction with long insulated conductors, such as overhead lines, can include a "bow wave" effect in which the inducing wave propagates along the line synchronously with the induced current wave, building up very large induced currents. The field produced by lightning decreases as  $1/r$  with distance, r, from the source, so that the bow-wave effect is much less prominent for lightning than it is for HEMP.

Perhaps the most important difference between lightning and the HEMP is their area of coverage. Lightning strikes one point in a large system such as a continental communication network, while the HEMP excites the entire network almost simultaneously. Large networks have been designed to cope with single-point outages, such as those that may occasionally occur because of lightning. We have no experience to assist us in determining the effect of a large number of simultaneous outages that might accompany HEMP, and it is virtually impossible to test hypotheses of system reactions with network-scale experiments. Furthermore, the system is not exposed to the HEMP during peacetime; we get no feedback from a "protected" system on the effectiveness of the protection. Thus, protecting large networks from the HEMP usually involves conservative protection of individual parts of the network in the hope that network hardness will follow from component hardness.

**10.3 HEMP INTERACTION WITH SYSTEMS.** HEMP interaction with systems may be separated into long-line effects and local effects. Long-line effects are the currents and voltages induced on long power lines, communication cable links, or even other conductors, such as pipelines. Some of these HEMP effects may be induced far away and guided to the facility along the conductor. Local effects are the currents and voltages induced directly on the facility shield, building structure, wiring, equipment cabinets, etc. These local effects are very difficult to evaluate analytically because of the complexity of the facility structure, the lack of information on the broadband electrical properties of many of the structural materials, and the extremely large number of interaction paths, facility states, and other complicating factors (10-2), (10-3). On the other hand, the local interactions can be evaluated experimentally with simulated HEMP fields that envelop the facility. The full length of the long lines connected to a facility can rarely be illuminated with simulated HEMP fields; the HEMP interaction with the long lines must usually be estimated analytically and simulated as an external excitation driving at the long-line port.

### 10.3.1 Current in Long Lines.

#### 10.3.1.1 Long Overhead Lines.

The currents induced on long straight overhead lines parallel to the Earth's surface by HEMP-like events have been analyzed thoroughly (10-4), (10-5), (10-6). If the line is over a perfectly conducting ground plane, the current has a waveform similar to the HEMP early-time waveform, except for a slightly longer risetime for lines more than a few feet high. For imperfectly conducting ground, such as soil, the imperfect reflection of the wave from the ground allows the line to be driven more strongly and for a longer time than if the ground were a good conductor.

The short-circuit current induced in a semi-infinite line (one extending from the observer to infinity) over soil for an exponential pulse of incident field is shown in Figure 10-4. The current is shown for horizontal polarization (dashed line) and vertical polarization (solid line) of the incident field. The curve  $\sigma = \infty$  is the current that would be induced in a wire over a perfectly conducting ground; this current is proportional to line height ( $h$ ), decay time constant ( $\tau$ ), and incident field strength ( $E_i$ ). The current in Figure 10-4 is normalized by containing the characteristic impedance of the line, the peak field ( $E_0$ ) and decay time constant ( $\tau$ ) of the incident field, the speed of light ( $c$ ), and a directivity function ( $D$ ). The directivity function ( $D$ ) depends on the azimuth angle ( $\phi$ ) between the wire and the vertical plane containing the Poynting vector of the incident wave, and on the elevation angle ( $\psi$ ) of the Poynting vector of the incident wave. The correction for finitely conducting ground is proportional to the incident field strength, the  $3/2$  power of the decay time constant, and the inverse square root of the soil conductivity ( $\sigma$ ).

For a  $300\Omega$  line 10 meters (33 feet) above soil having  $10^{-3}$  siemens/meter (S/m) conductivity, an incident 15 kV/m exponential pulse with 250 nanoseconds (ns) decay time-constant will induce a short-circuit current of about 10kA on the line. Vertically polarized waves induce larger currents than horizontally polarized waves, but in the latitudes of the mainland United States, the HEMP fields are predominantly horizontally polarized. Thus, only 15 kV/m was used in this example, even though the peak HEMP field may be much larger than 15 kV/m. More sophisticated analyses that take into account the burst point, the observer point, and their effect on HEMP polarization and waveform give peak short-circuit currents between 5 and 10 kA for the early-time HEMP. The open-circuit voltage induced at the end of the semi-infinite line is the product of the short-circuit current and the characteristic impedance ( $Z_0$ ). For the  $300\Omega$  line in this example, the open-circuit voltage would be 3 megavolts (MV).

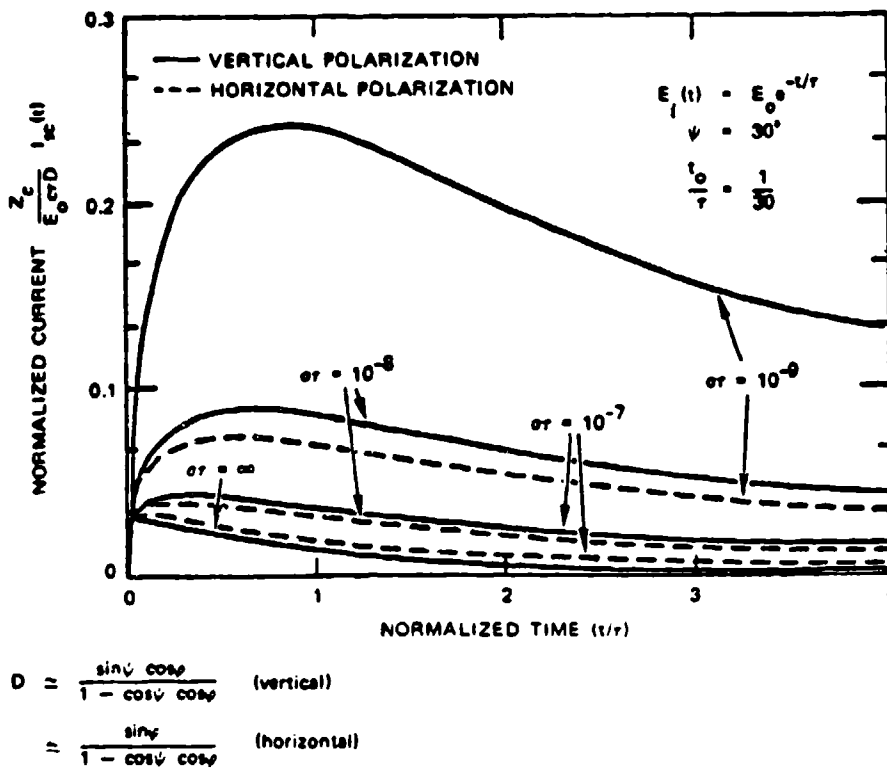


Figure 10-4. Short-Circuit Current Induced at the End of a Semi-Infinite Above-Ground Wire by an Exponential Pulse

#### 10.3.1.2 Long Buried Lines.

As noted in 10.3.1.1, imperfectly conducting soil does not completely reflect the incident field; some of the incident wave is transmitted into the soil. This field in the soil can induce current in underground cables, pipes, and other conductors. However, because the velocity of propagation of a wave is much less in soil than in air, the bow-wave effect is almost negligible on buried conductors. Furthermore, the attenuation on buried conductors is greater than on overhead lines because of the proximity of the lossy soil to the buried conductor. For conductors in contact with the soil (i.e., buried bare conductor), the current at any observation point is determined primarily by coupling within one skin-depth of the observation point. Current induced at points farther away is so strongly attenuated by the soil that it adds little to the total current at the observation point.

The current induced in a long buried cable is shown in Figure 10-5 for various depths of burial, as given by a depth parameter  $\tau_d = \mu \sigma d^2$ , where  $d$  is the depth of burial. The current is normalized to the inductance per unit length ( $L$ ) of the cable, the magnitude of the incident exponential pulse ( $E_0$ ), the decay time-constant ( $\tau$ ), the soil time-constant ( $\tau_0 = c/\sigma$ ), and a directivity function ( $D$ ). The induced current is proportional to the incident field strength ( $E_0$ ), the  $3/2$  power of the decay time-constant ( $\tau$ ), and the inverse square root of the soil conductivity ( $\sigma$ ). For a horizontally polarized, vertically incident pulse having  $E_0 = 50$  kV/m and  $\tau = 250$  ns, a long cable buried near the surface ( $d = 0$ ) in soil of conductivity  $10^{-3}$  S/m will have about 2.8 kA induced in it.

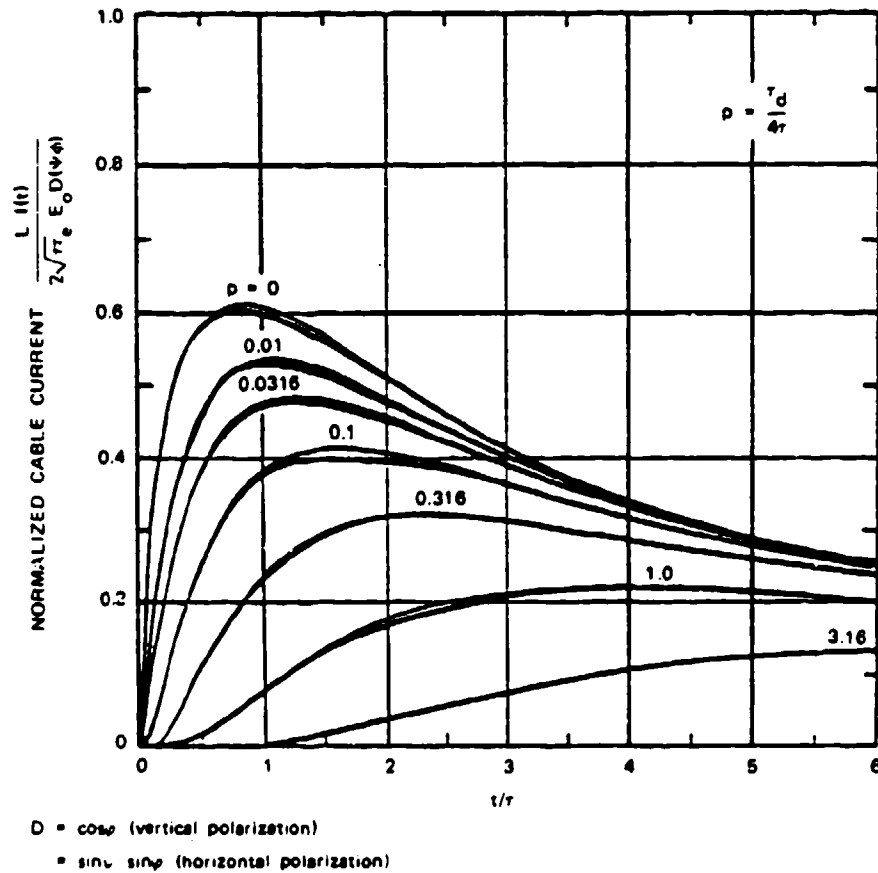


Figure 10-5. The Normalized Current Waveform for Various Values of the Depth Parameter  $p$  (Exponential Pulse)

**10.3.1.3 Vertical Structures.** The HEMP interacts with vertical structures, such as radio towers, waveguides, and cables to overhead antennas, and downloads from power and communication lines in much the same manner as it interacts with horizontal lines, except that it is the vertical component of the electric field that drives the vertical structures. The current induced in a download from an overhead power line is shown in Figure 10-6. Because the line is short and the angle of incidence is only  $30^\circ$ , little bow-wave effect is observable. The peak current is also limited by the line height in this example; for taller structures, the leading edge of the current wave will continue to rise as the integral of the incident wave. The current will increase with structure height for structures up to a few hundred feet high.

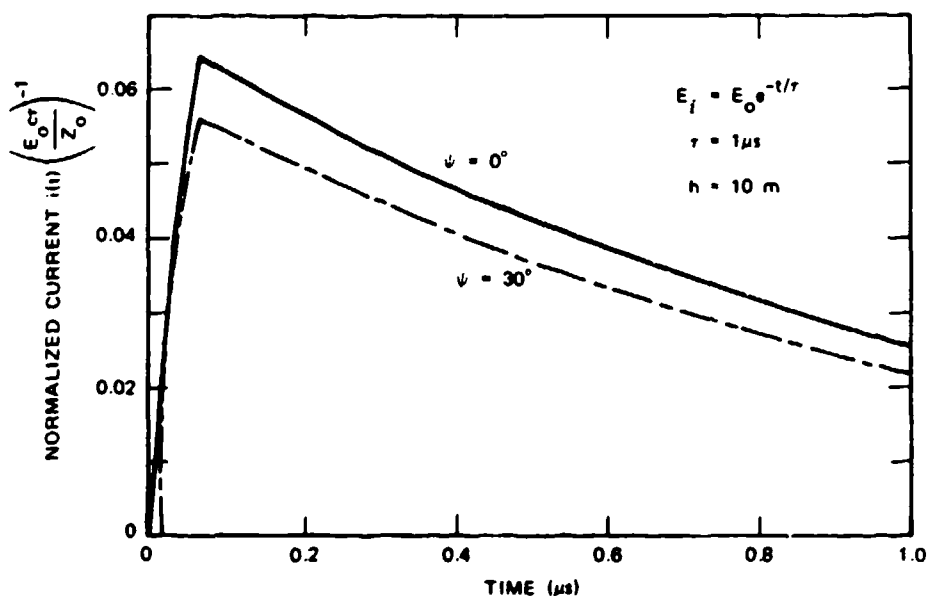


Figure 10-6. Short-Circuit Current Induced at the Base of a Vertical Riser by a Vertically Polarized Incident Wave

### 10.3.2 HEMP Interaction with Local Structure.

**10.3.2.1 Shields.** The HEMP fields incident on a closed shield induce surface currents and charge displacements on the outer surfaces of the shield. If the shield is continuous metal (i.e., it has no opening or discontinuities in its surface) and about 1 mm thick, the voltage induced in circuits inside the shield by the HEMP will be very small. Table 10-1 shows the voltage induced in the largest single-turn loop that can be placed inside a spherical shield of 10 meters radius by a zero-rise-time 50 kV/m incident exponential pulse having a 250 ns decay time-constant ( $10^{-7}$ ). Note that even for a shield as thin as 0.2 mm, the induced voltage is less than 1 V; shields made of workable thicknesses of common metals do not allow significant HEMP interaction with internal circuits. Possible exceptions to this conclusion are those shields that are long and of small cross section, such as the shields on intrasite cables.



Table 10-1

## Shielding by Diffusion

Shield Thickness (mm)	Internal Voltage Induced in Loop *		
	Copper $5.8 \times 10^7$ S/m	Aluminum $3.7 \times 10^7$ S/m	Steel $6 \times 10^6$ S/m ( $\mu_r = 200$ )
0.2	0.34 V	0.85 V	0.076 V
1.0	2.6 mV	6.4 mV	1.1 mV
5.0	21.0 $\mu$ V	51.0 $\mu$ V	15.0 $\mu$ V

\*Peak voltage induced in a loop by radius 10 m inside a spherical shield or radius 10 m illuminated by a high-altitude EMP (by diffusion through walls only).

**10.3.2.2 Penetrating Conductors.** Conductors, such as power and signal wires, that pass through the shield, as illustrated in Figure 10-7, may allow very large currents and voltages to be delivered to internal circuits. The current on the wire just inside the shield is about equal to the current just outside the shield; the wire is a 0 dB compromise of the shield. (At high frequencies, the capacitance between the wire and the shield wall may cause some attenuation of the wire current, but this effect is negligible at frequencies such that  $1/j\omega C \ll Z_c$ , where  $C$  is the wire-to-shield capacitance and  $Z_c$  is the load impedance on the wire. (For nominal values of  $C = 10$  pF and  $Z_c = 200\Omega$ ,  $f = 80$  MHz for a 6-dB loss.) Thus a major concern for HEMP interaction is the penetrating conductor that can guide HEMP-induced waves through shield walls. As discussed above and illustrated in Figure 10-7, the shield is effective in excluding the incident electromagnetic waves, but it has little effect on the waves guided through it on insulated penetrating conductors (10-8).

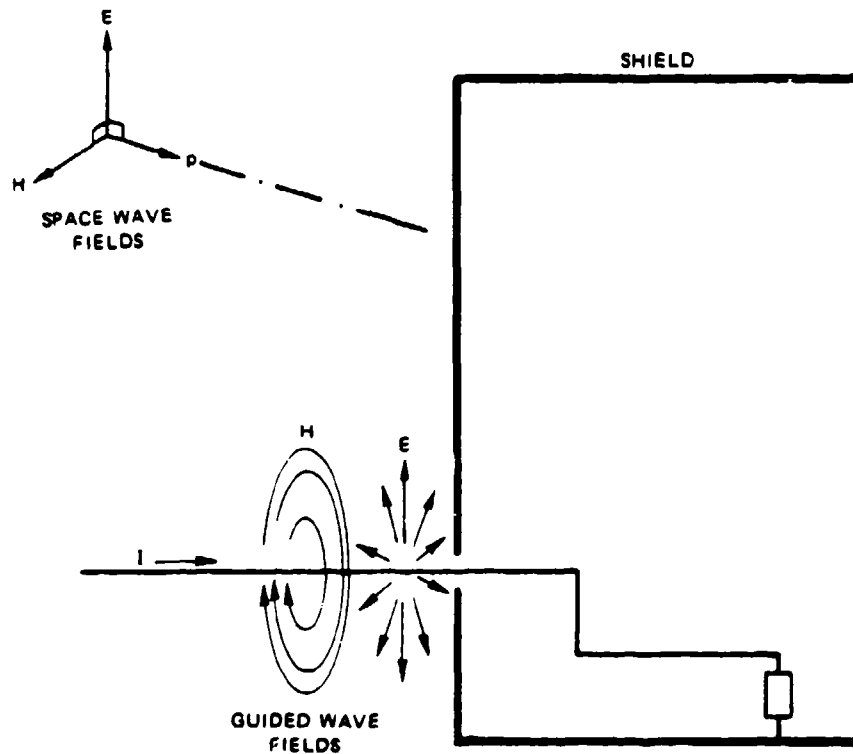
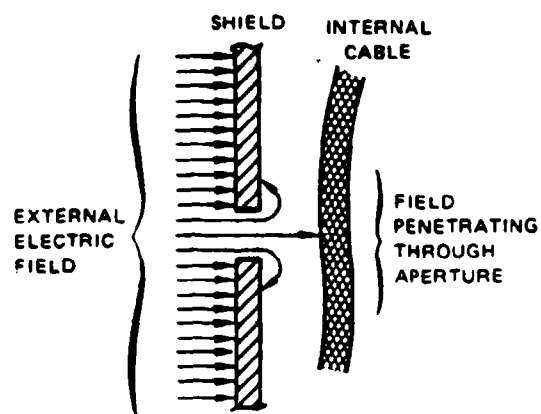


Figure 10-7. Shield to Exclude Electromagnetic Fields

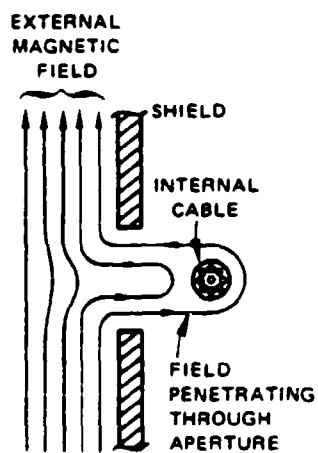
### 10.3.2.3 Apertures.

Apertures in the shield surface allow the external HEMP-induced fields to penetrate through the shield and interact with internal wiring or other conductors, as illustrated in Figure 10-8. The external electric field ( $E_n$ ) associated with the surface charge density ( $q = \epsilon_0 E_n$ ) can induce charge on internal cables as illustrated in Figure 10-8a. The external magnetic field ( $H_t$ ) which has the same magnitude as the surface current density ( $J$ ) can penetrate through the aperture to link internal circuits, as illustrated in Figure 10-8b. Since the current induced by the electric field is proportional to  $dE_n/dt$  and the voltage induced by the magnetic field is proportional to  $dH_t/dt$ , the aperture coupling emphasizes the fast-changing parts, or high-frequency spectrum, of the HEMP-induced transient. However, it is important to recognize that it is in the rate of rise that the HEMP stress is dominant over the other external sources.

The maximum open-circuit voltage induced by a rate of change of external magnetic field of  $10^{11} \text{ Am}^{-1}\text{s}^{-1}$  penetrating a circular aperture 5 cm in radius is over 600 V. This rate of change of the field is characteristic of the HEMP, and the analysis leading to the induced voltage is based on a wire directly across the aperture just inside the shield. Thus, aperture coupling is an important consideration in HEMP interaction analysis. Apertures in facility shields take many forms; they range from open doors and windows to the discontinuities at riveted or bolted joints in the shield.



(a) ELECTRIC FIELD



(b) MAGNETIC FIELD

Figure 10-8. Electromagnetic Penetration Through Small Apertures

#### 10.4 PROTECTION AGAINST HEMP.

There are important considerations in designing this protection that affect the value that can be placed on HEMP protection. The HEMP protection adds cost to the facility, and the value received for the added cost is confidence that the facility will survive HEMP. This implies that (1) the protection against HEMP can be verified, and (2) this protection is retained and can be maintained throughout the life of the facility. The protection has low value when it is designed in such a way that it is difficult to verify or maintain. The protection may be difficult to verify when the HEMP-induced stresses inside the facility are large enough to cause spurious arcing or other insulation breakdown. It may also be difficult to verify when it depends on unknown or uncontrolled electromagnetic properties of materials used in the facility. Finally, hardness verification will be difficult if the number of features that must be tested is very large. For example, if the HEMP-induced stress is large deep inside the facility, the number of system states, modes of excitation, stress waveforms, etc., that must be evaluated may be enormous.

Since HEMP does not ordinarily occur during peacetime, degradation of the protection is not evident from peacetime operation of the facility. Therefore, the HEMP protection has greatest value if it is durable. The protection should not be degraded by normal use and maintenance of the facility. The protection should not depend on extraordinary configuration control. It must accommodate facility growth and modification. Components critical to the protection should be few in number, accessible, and testable.

Protecting communication facilities against the HEMP typically consists of developing a closed HEMP barrier about the facility. The barrier consists of a shield to exclude the incident space waves and various barrier elements on the essential penetrating conductors and in the apertures required for personnel and equipment. The number of penetrating wires, apertures, and other features that must be evaluated to verify the HEMP protection is kept as small as possible. In addition, attention is given to the number of system states or configurations for which the protection must be determined. Durability and accessibility of the protection elements are also important.

##### 10.4.1 HEMP Barrier.

10.4.1.1 Shield. The facility-level shield used for protection against HEMP is typically fabricated from welded sheet steel. The thickness is usually selected for ease of fabrication, but in areas where exceptional mechanical abuse is likely, mechanical strength, as well as workability, may be a consideration. Shield assembly is typically accomplished by continuous welding, brazing, hard soldering, or other fused-metal process to minimize the number of discontinuities in the shield (a weld or other fused-metal joint is considered continuous metal).

##### 10.4.1.2 Penetrating Conductors.

Concepts for penetrating conductor treatment are illustrated in Figure 10-9. Penetrating conductors that can be grounded, such as plumbing, waveguides, grounding cables, and cable shields, are bonded to the shield wall at their point of entry by peripherally welding them to the wall or by the use of clamps, collets, etc., that peripherally bond the penetrating conductor to the shield with little or no discontinuity.

Signal and power wires that need not penetrate the shield should not penetrate the shield. Wires that must penetrate the shield must be treated with a barrier element, such as a filter or surge arrester, that closes the barrier above a voltage threshold or outside the passband required for signal or power transmission.

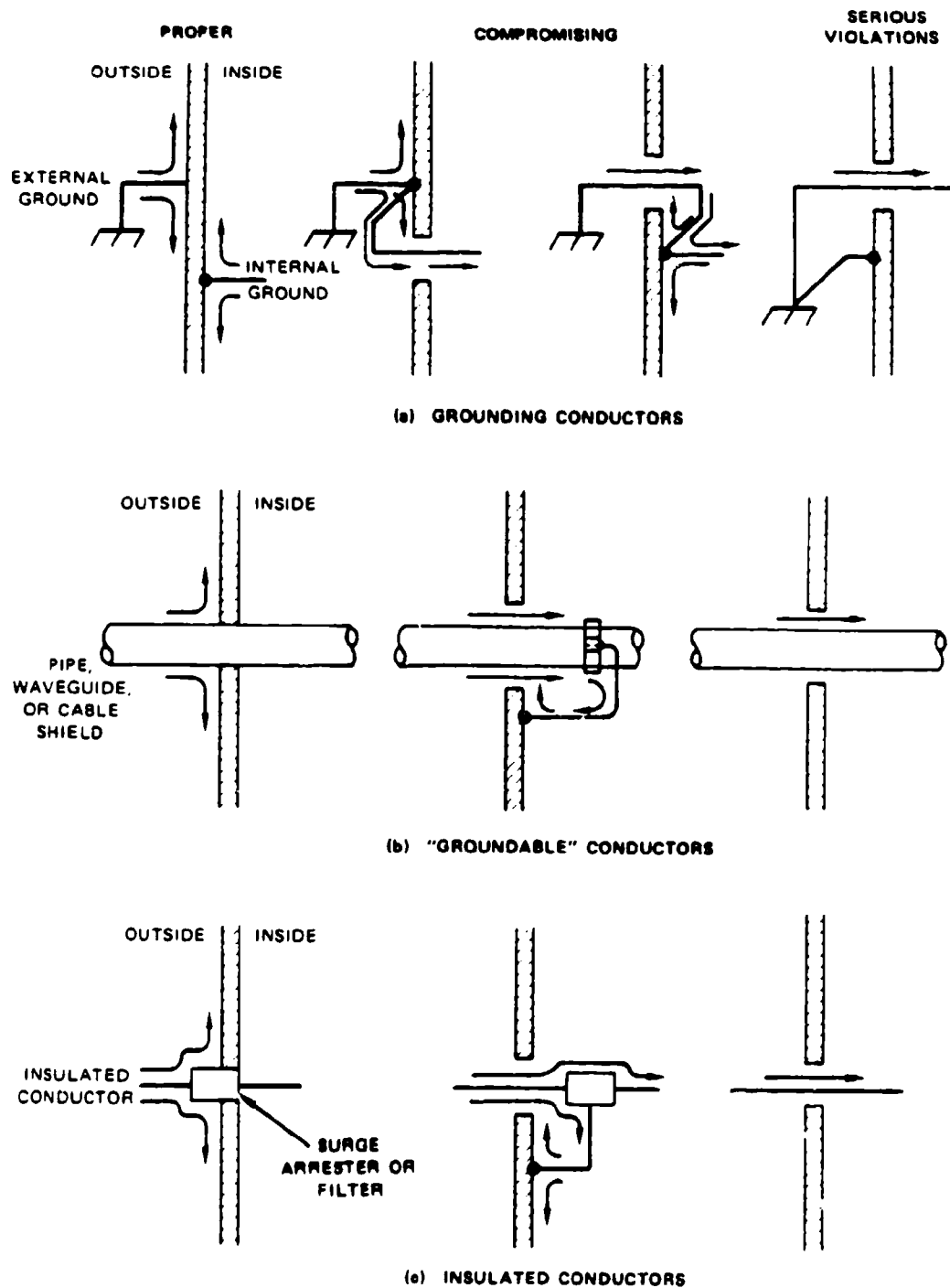


Figure 10-9. Shielding Integrity Near Interference-Carrying External Conductors

**10.4.1.3 Apertures.** No unnecessary openings or discontinuities in the shield should be allowed. Those openings necessary for personnel and equipment loading and for ventilation should be designed to limit electromagnetic field penetration. Openings that must permit air flow or light passage can be made more opaque to HEMP waves by covering them with mesh or, preferably, honeycomb waveguide-beyond-cutoff, as illustrated in Figure 10-10. High traffic entryways can use waveguide-beyond-cutoff tunnels with doors at each end (without the doors, the highest frequencies in the HEMP spectrum can penetrate through a tunnel large enough for personnel to walk through). Where possible, discontinuities in the shield should be eliminated by continuous welding or a similar process. Those that are necessary for equipment installation and maintenance should be electromagnetically sealed with durable bonding techniques, such as resilient RFI gaskets and small bolt spacing. Where access is required infrequently, it might be practical to weld the equipment entry door shut; large cargo doors are large compromises to shielding integrity and difficult to seal effectively and durably.

#### **10.4.2 Allocation of Protection.**

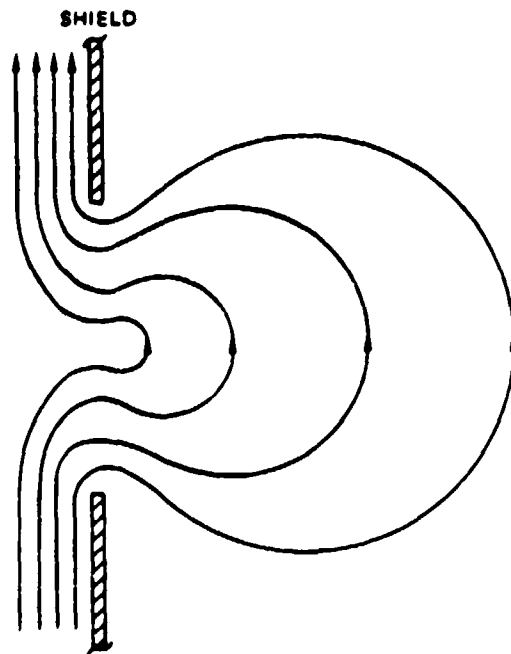
##### **10.4.2.1 Amount of Protection Needed.**

The amount of protection needed depends to some extent on how failure is defined for the system. For communications facilities, the threshold for failure, or the minimum acceptable performance, may be defined by a maximum allowable outage time or error rate. In some cases, the principal requirement is that the system not damage itself so that it can be restarted and restored to service after an attack involving HEMP. The definition of system failure, or operating requirements, should be prescribed in the system specification; it will be determined by many factors in addition to HEMP.

In determining the amount of HEMP protection required, it is important to be able to define a transient tolerance or susceptibility level for the facility or the equipment in the facility. Since most communications equipment have no transient "withstand" requirement, except perhaps on the power terminals, we cannot obtain the required tolerance from the equipment specification. Nevertheless, it is possible to define a transient stress at or below which the equipment performance will be unaffected. Zero stress certainly satisfies this definition, but more practical values can be found. For example, the equipment tolerates its operating signal levels, and it tolerates the peacetime transient stress inside the facility. Neither of these is a trivial value of stress, and we can be assured that if the HEMP-induced stress is made small compared to either, the presence of HEMP will not cause the equipment to malfunction. Additional information on transient withstand requirements may be found in MIL-STD-461C.

The equipment or internal circuit threshold defined in terms of known peacetime tolerances has several advantages: (10-9)

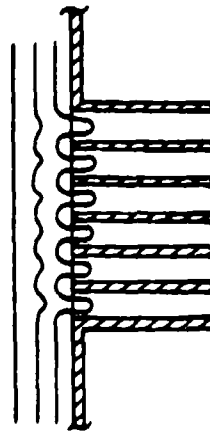
- (1) It takes advantage of known equipment "withstand" capability; no more HEMP protection is necessary than that required to reduce HEMP transients to a safe margin below this known tolerance.
- (2) It is not necessary to determine the HEMP response of circuits and structural elements inside the equipment; this greatly reduces the complication of hardness verification and maintenance.
- (3) It is possible to place all HEMP requirements at the facility barrier, so that concern for interior configuration control and internal states are alleviated.



(a) SINGLE APERTURE



(b) MANY SMALL APERTURES



(c) ARRAY OF WAVEGUIDES  
BEYOND CUTOFF

Figure 10-10. Magnetic Field Penetration of Apertures

#### 10.4.2.2 Where Protection is Applied.

As noted above, HEMP protection must be designed to accommodate hardness verification procedures. The most easily verified protection requires the least number of tests and the least number of assumptions to establish the integrity of the protection. For example, suppose the facility is a node with a 100-pair cable linking it to other parts of a network. Because of unbalanced and nonlinear terminations, there may be  $200^2$  two-wire stresses and susceptibilities to evaluate at the cable penetration of the facility barrier. Inside the facility, the 200-wire cable may branch into 1000 or more wires and equipment terminals. Thus if the HEMP stress is allowed to be dominant (larger than known peacetime stresses) inside the barrier,  $1000^2$  transient stresses and susceptibilities must be evaluated (or assumed unimportant). In addition, in the latter case, all the internal interactions between the 1000 wires and other internal circuits must be assessed (or assumed unimportant). Typically, both the number of features to be evaluated and the number of assumptions necessary increase with the depth into the system at which hardness verification is attempted.

Therefore, for facilities whose protection against HEMP has high value (i.e., where confidence in the protection is important), the protection is placed at the system-level barrier, and the protection at this level is sufficient that the HEMP-induced stress is not dominant inside this barrier.

10.4.2.3 Terminal Protection Devices. Problems from HEMP are expected to arise from the antennas and connecting cables, long interconnecting leads and cables between equipments, and the ac power lines. Antennas, connecting cables, and the front-end of the associated communications equipments in particular will be subjected to very large voltages and currents. The protective technique or device must protect the equipment without adversely affecting its performance, and must be capable of withstanding the effects of both EMP-induced transients and other transients in the system. The latter two considerations may severely limit applications of many of the protective devices at rf unless they are modified or used in conjunction with other components.

10.4.2.3.1 Spark Gaps and Gas Tubes. Spark gaps are one of the oldest forms of surge arrester. A spark gap is a pair of electrodes, insulated by air or other gas, spaced so that the gap will break down when the voltage exceeds a specified level. The insulating gas pressure varies from a fraction of an atmosphere to several atmospheres, and the electrode spacing varies from a few millimeters in carbon blocks to several inches in large lightning arresters used for power equipment. Firing voltages range from about 1 kV for some carbon blocks to hundreds of kV for large lightning arresters. Large spark gaps can handle large charge transfers (many coulombs). In the nonconducting state, spark gaps behave as open circuits or small capacitances. The spark-gap firing voltage increases with the rate-of-rise of the applied voltage. Thus, for the large rates-of-rise encountered in EMP-induced voltages, the firing voltage may be several times as large as the rated static firing voltage. When spark gaps are used on energized lines, some provision must be made to assure that the discharge will be extinguished. Frequently, a metal-oxide varistor (MOV) is used in series with the spark gap to ensure arc extinction after the surge.

Gas tubes are spark gaps with a low-pressure gas so that lower firing voltages can be achieved. Firing voltages below 100 V are available for commercial gas tubes. The gas tubes are generally more limited in their peak current and charge transfer capability than the spark gaps. Gas tubes are used primarily for secondary protection of wire pairs entering a facility from a long external shielded cable, or for exposed intrafacility



wiring. Balanced two-wire models are available that allow ionization from the first discharge to cause immediate conduction of both halves of the tube so that circuit imbalance is minimized. Coaxial models are also available for use on coaxial lines such as antenna feed cables. Gas tubes have small capacitances and virtually no loss in the nonconducting states. The glow state occurs in circuits whose impedance limits the discharge current to less than about 100 mA; the voltage across the tube in this state is about 100 V. The arc state occurs when large currents are caused to flow; the voltage across the tube in the arc state is usually 10 to 20 V. Gas tubes should not be used on energized lines that can sustain the arc or glow discharge.

Spark gaps and gas tubes display a negative dynamic resistance at the firing point, where a decrease in voltage across the device is accompanied by an increase in current through it. This property of spark gaps and gas tubes sometimes leads to unpredicted instabilities in the protected circuits. In addition, the discharge is a sudden change in voltage and current that may shock-excite the protected circuit. It is usually recommended that a linear filter be placed between the device and the protected circuits to minimize the effects of the negative dynamic resistance and shock excitation.

**10.4.2.3.2 Metal-Oxide Varistors.** MOVs are capable of diverting currents up to tens of kiloamperes and, when packaged and installed to minimize terminal and lead inductance, they are effective for large rate-of-rise transients. Although they are nonlinear, MOVs do not display the negative dynamic resistance and shock excitation characteristics of the spark gaps and gas tubes. Their nonlinearity may produce intermodulation effects in RF circuits. The MOV stops conducting when the applied voltage decreases below the "knee" of the V-I curve. It is ideal for protecting energized lines, since it has no current-extinguishing problems. The MOV typically adds nanofarads of shunt capacitance and megohms of shunt resistance to the protected circuit. It should be used with caution on high-frequency circuits and high-impedance circuits. The maximum energy dissipation capability for large MOVs is tens of kilojoules. Just above the failure threshold, they usually fail as a short circuit or low resistance. However, for energies well above the failure threshold, the devices may be physically destroyed, sometimes explosively.

**10.4.2.3.3 Semiconductors.** A number of avalanche devices are available for use as surge limiters. The semiconductor devices limit at lower voltages (1 to 100 V) than the MOVs and gas tubes, but they are less tolerant of large peak currents and large energies than the other devices. Peak current ratings up to about 100 A are available. Because the devices themselves may be damaged by transients arriving on external wires and cables, they are not recommended for facility-level use. They may be used to protect equipment inside the facility and circuits that are entirely inside the shielded facility. The semiconductor devices add nanofarads of shunt capacitance to the protected circuit and may aggravate intermodulation problems.

**10.4.2.3.4 Filters.** Linear filters may also be used as barrier elements on penetrating wires, but at the outer (facility-level) barrier, filters are always used in combination with surge arresters. On power lines, for example, the line filter usually cannot tolerate the peak voltages, so a spark-gap surge arrester is used to limit the voltage, and the filter isolates the interior circuits from the negative dynamic resistance and shock excitation of the spark-gap discharge. The shunt input capacitance of the filter may also be used to reduce the rate-of-rise of the voltage, so that the firing voltage of the surge arrester will be lower. A variety of low-pass, bandpass, and high-pass filters is available for power and signal line protection.

#### 10.4.2.4 Waveguide Penetration of Facility Shield.

##### 10.4.2.4.1 Introduction.

Waveguides, like other external conductors that penetrate the facility shield, can allow transients to propagate into the facility if they are not made continuous with the shield in the manner illustrated in Figure 10-11. Ideally, the waveguide wall should make continuous contact with the facility shield around the entire periphery or the waveguide combination. All of the waveguide current would then flow onto the outer surface of the facility shield; the external transients could only penetrate to the interior by diffusion through the waveguide or facility wall.

In practice this continuous peripheral contact between the waveguide and the shield can be achieved by welding or soldering the waveguide to the wall. Two ways of implementing this connection are illustrated in Figure 10-12, where waveguide feedthrough sections are installed in the facility shield wall (or in a panel that is welded or bolted to the wall). In both cases, a waveguide section with two flange joints is used to allow the internal waveguide signal to pass through the wall but keep the external transient interference outside the facility. This method of treating the waveguide allows some flexibility in the waveguide plumbing inside and outside the facility, since only the feedthrough section is permanently attached to the wall.

For microwave receiving systems operating with very small signals, the fraction of a dB loss in the joints and the possibility of additional loss from distortion about the weld may be undesirable, although the weld distortions can be eliminated by machining or reforming the welding operation. Where these losses are intolerable, some alternate methods of attaching the waveguide to the facility shield are available. In the following sections, two of these methods are described. Although these methods can be used satisfactorily, they are generally less rugged and more susceptible to corrosion and other degradation than the welded feedthrough sections of Figure 10-12.

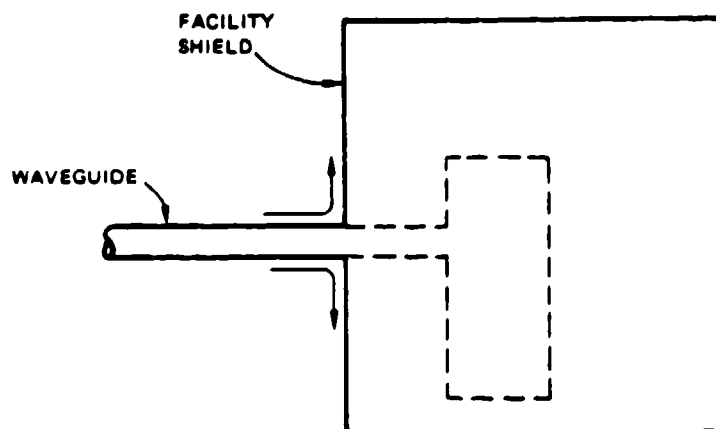


Figure 10-11. Exclusion of Waveguide Current from Interior of Facility

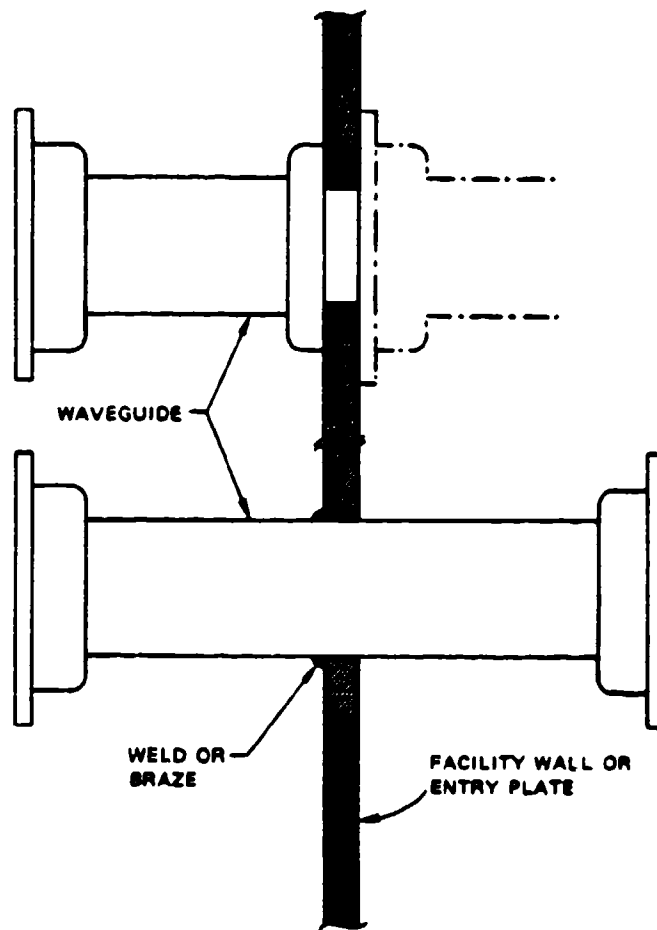


Figure 10-12. Waveguide Feedthroughs

**10.4.2.4.2 In-Line Waveguide Attachment.** Connecting the waveguide to the shield without the feedthrough section and flange joints requires an in-line connection. In-line connections are somewhat inconvenient, because the waveguide penetration hole in the facility must be fairly accurately located so that it is aligned with the waveguide ports on the internal equipment and the external plumbing. In addition, the hole in the wall must be large enough to pass a waveguide flange, yet must be effectively closed by the attachment mechanism. Finally, the attachment to the waveguide must accommodate misalignment of the waveguide with the axis of the hole. If we further prohibit welding or brazing because of the potential distortion and damage to the internal finish of the waveguide, we are limited to soft soldering, mercury wetting, and clamping to make the electrical connection to the waveguide. Because of its environmental problems and its tendency to dissolve waveguide materials, mercury wetting has not been proposed to make the connection. The use of soft solder bonds also is prohibited by MIL-STD-188-124A on conductive paths subject to lightning or power fault currents. The following procedures are acceptable for bonding waveguides or cables to a designated RF shield, barrier or entrance plate.

**10.4.2.4.2.1 Sleeve and Bellows Attachment.** In this method, illustrated in Figure 10-13, the connection to the waveguide is made with a snug-fitting sleeve over the waveguide. The sleeve may be installed on the guide before the end fittings are installed, or a split sleeve may be used so that it can be installed at any time. The preferred method of attaching the sleeve to the waveguide is to soft solder the sleeve to the guide with a eutectic lead-tin alloy. For split sleeves, however, it will probably be necessary to provide mechanical support with a clamp, as illustrated in the figure. If even eutectic soldering cannot be tolerated, a clamping alone may suffice, if the sleeve is slitted to allow it to grip the waveguide and if the sleeve and waveguide are both clean and protected so that they remain clean. To help prevent distortion of the waveguide by the clamp, it is recommended that a neoprene or other resilient cushion be used between the clamp and the sleeve. The flange on the sleeve and the bellows and its flanges can be welded together without damaging the waveguide. Details of their design are optional, but the bellows and flanges must be large enough to pass the waveguide flange if the bellows assembly is to be installed in the field after the waveguide is assembled.

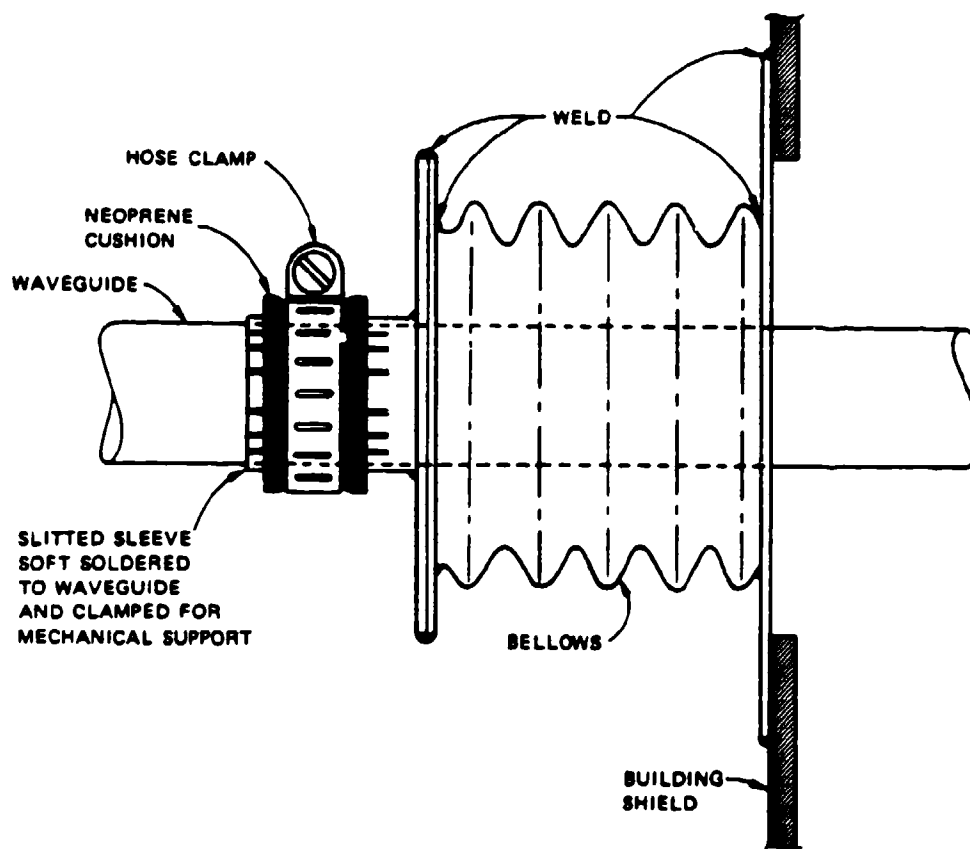


Figure 10-13. Bellows with Slitted Sleeve Waveguide Attachment

#### 10.4.2.4.2.2 Braided Wire Sleeve.

A somewhat less effective, but usually adequate attachment to the waveguide can be made with a braided wire sleeve. As illustrated in Figure 10-14, the braided wire sleeve is necked down and soldered to the waveguide and flared out over a collar on the facility shield wall, where it is also soldered or welded. For mechanical strength, both of these attachments should be reinforced with a hose clamp and cushion, as was used with the rigid sleeve. And as with the rigid sleeve, the clamp may be used without solder at the waveguide if soldering cannot be tolerated, but, as before, both the braid and the waveguide must be clean when assembled and remain clean after assembly.

The braided wire sleeve must expand into a large enough hoop to enable the waveguide end fittings to pass through (unless the sleeve is installed before the fittings are installed). In addition, it is desirable that the sleeve have an optical coverage of at least 85%. Thus, the sleeve design is fairly stringent because large expansion is usually accompanied by low coverage.

For both the bellows and the braided wire sleeve attachments, it is recommended that the attachment mechanism be placed inside the facility wall and that a weatherproof boot or other external seal be installed to keep moisture and other foreign matter out of the attachment.

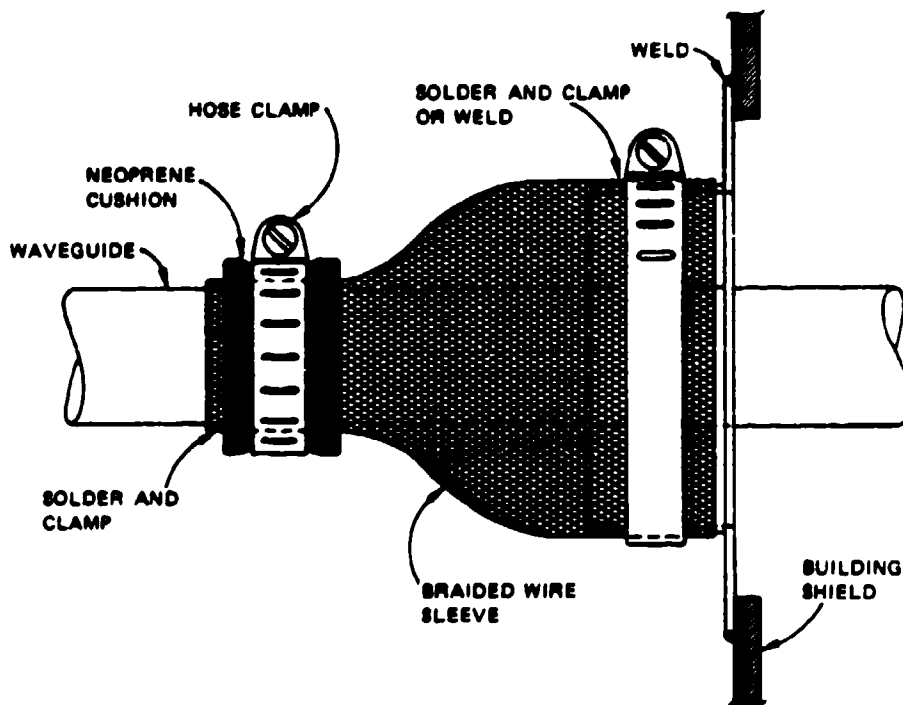
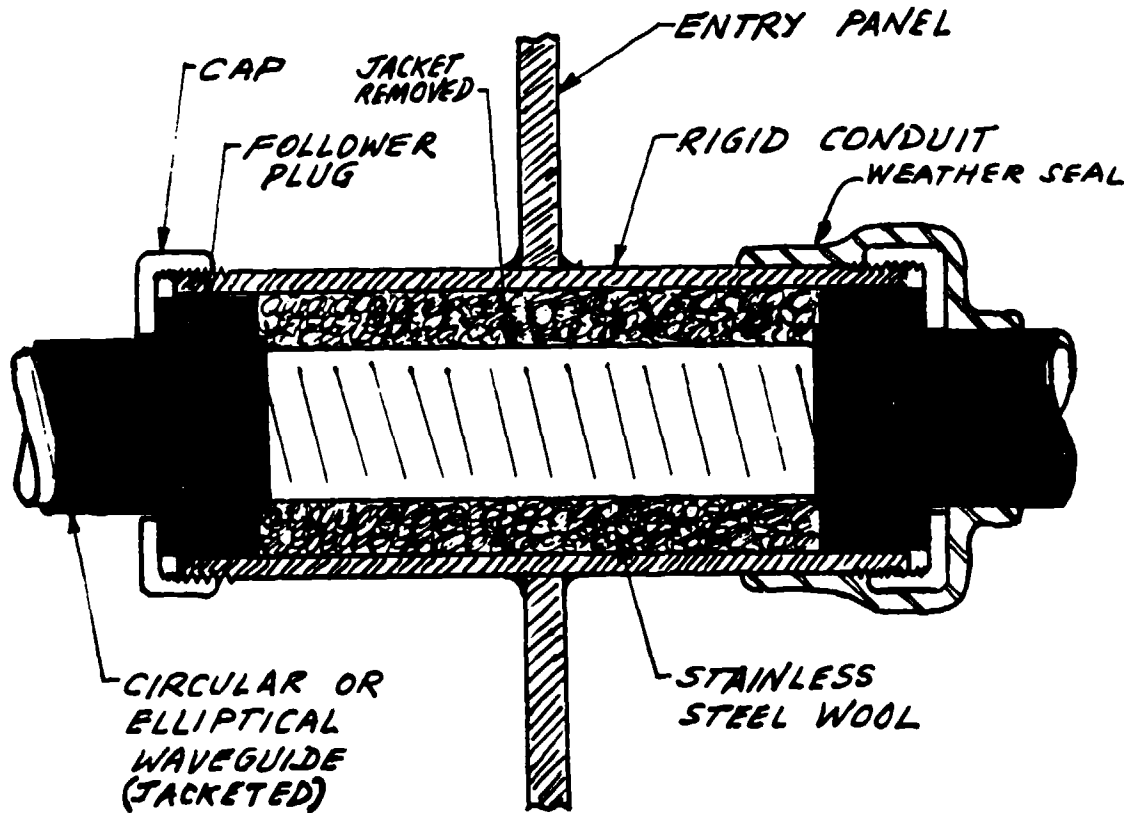


Figure 10-14. Braided Wire Sleeve Clamped to Waveguide

**10.4.2.4.2.3 Stuffing Tube for Waveguide.** In this method, illustrated in figure 10-15, the connection to the waveguide is made with a highly compressed stainless steel wool placed between rigid conduit and the bare waveguide. The conduit must be installed over the waveguide before the end fittings are installed. The follower plugs serve to compress the steel wool and also aid in weatherproofing and protecting the bond from corrosion. After all weather proofing has been completed, the rigid conduit should be bonded to the entry panel or facility shield by welding or brazing.



**NOTES:**

1. Steel wool lightly compressed by follower plugs.
2. Steel wool and waveguide surface must be protected against corrosion.

Figure 10-15. Stuffing Tube for Waveguide

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CHAPTER 11

NOTES

**11.1 SUBJECT TERM (KEY WORD) LISTING.** Key words contained in Volume I and Volume II of this handbook include:

Grounding  
Bonding  
Shielding  
Facility Ground System  
Lightning Protection Subsystem  
Fault Protection Subsystem  
Signal Reference Subsystem  
Earth Electrode Subsystem  
Single-Point Grounding  
Multipoint Grounding  
Equipotential Ground Plane  
Air Terminal (Lightning Rod)  
Electromagnetic Pulse (EMP)  
Lower Frequency Ground  
Higher Frequency Ground  
Phase Conductor  
Grounding Conductor (Green Wire)  
Grounded Conductor  
Neutral Conductor  
Ring Ground  
Fall-of-Potential Ground Test  
Shielding Effectiveness  
Zone (Cone) of Protection  
Power System Grounding  
Signal Grounding  
Facility Shielding  
Equipment Shielding  
Corrosion  
Down Conductor, Lightning  
Cathodic Protection

APPENDIX A

GLOSSARY

**ABSORPTION LOSS** -- The attenuation of an electromagnetic wave as it passes through a shield. This loss is primarily due to induced currents and the associated  $I^2R$  loss.

**AIR TERMINAL** -- The lightning rod or conductor placed on or above a building, structure, tower, or external conductors for the purpose of intercepting lightning.

**APERTURE** -- An opening in a shield through which electromagnetic energy passes.

**BALANCED LINE** -- A line or circuit using two conductors instead of one conductor and ground (common conductor). The two sides of the line are symmetrical with respect to ground. Line potentials to ground and line currents are equal but of opposite phase at corresponding points along the line.

**BOND** -- The electrical connection between two metallic surfaces established to provide a low resistance path between them.

**BOND, DIRECT** -- An electrical connection utilizing continuous metal-to-metal contact between the members being joined.

**BOND, INDIRECT** -- An electrical connection employing an intermediate electrical conductor or jumper between the bonded members.

**BOND, PERMANENT** -- A bond not expected to require disassembly for operational or maintenance purposes.

**BOND, SEMIPERMANENT** -- Bonds expected to require periodic disassembly for maintenance, or system modification, and that can be reassembled to continue to provide a low resistance interconnection.

**BONDING** -- The process of establishing the required degree of electrical continuity between the conductive surfaces of members to be joined.

**BUILDING** -- The fixed or transportable structure which houses personnel and equipment and provides the degree of environmental protection required for reliable performance of the equipment housed within.

**CABINET** -- A protection housing or covering for two or more units or pieces of equipments. A cabinet may consist of an enclosed rack with hinged doors.

**CASE** -- A protective housing for a unit or piece of electrical or electronic equipment.

**CHASSIS** -- The metal structure that supports the electrical components which make up the unit or system.

**CIRCULAR MIL** -- A unit of area equal to the area of a circle whose diameter is one mil (1 mil = 0.001 inch). A circular mil is equal to  $\pi/4$  or 78.54 percent of a square mil (1 square mil =  $10^{-6}$  square inch). The area of a circle in circular mils is equal to the square of its diameter in mils.

**CIRCUIT** -- An electronic closed-loop path between two or more points used for signal transfer.

**COMMON-MODE VOLTAGE** -- That amount of voltage common to both input terminals of a device.

**COMMON-MODE REJECTION** -- The ability of a device to reject a signal which is common to both its input terminals.

**CONDUCTED INTERFERENCE** -- Undesired signals that enter or leave an equipment along a conductive path.

**COPPER CLAD STEEL** -- Steel with a coating of copper bonded on it.

**COUPLING** -- Energy transfer between circuits, equipments, or systems.

**COUPLING, CONDUCTED** -- Energy transfer through a conductor.

**COUPLING, FREE-SPACE** -- Energy transfer via electromagnetic fields not in a conductor.

**CUTOFF FREQUENCY** -- The frequency below which electromagnetic energy will not propagate in a waveguide.

**DEGRADATION** -- A decrease in the quality of a desired signal (i.e., decrease in the signal-to-noise ratio or an increase in distortion), or an undesired change in the operational performance of equipment as the result of interference.

**DOWN CONDUCTOR, LIGHTNING** -- The conductor connecting the air terminal or overhead ground wire to the earth electrode subsystem.

**EARTH ELECTRODE SUBSYSTEM** -- A network of electrically interconnected rods, plates, mats, or grids installed for the purpose of establishing a low resistance contact with earth.

**ELECTRIC FIELD** -- A vector field about a charged body. Its strength at any point is the force which would be exerted on a unit positive charge at that point.

**ELECTROMAGNETIC COMPATIBILITY (EMC)** -- The capability of equipments or systems to be operated in their intended operational environment, within designed levels of efficiency, without causing or receiving degradation due to unintentional EMI. EMC is the result of an engineering planning process applied during the life cycle of equipment. The process involves careful considerations of frequency allocation, design, procurement, production, site selection, installation, operation, and maintenance.

**ELECTROMAGNETIC INTERFERENCE (EMI)** -- Any electrical or electromagnetic phenomenon, manmade or natural, either radiated or conducted, that results in unintentional and undesirable responses from, or performance degradation or malfunction of, electronic equipment.

**ELECTROMAGNETIC PULSE (EMP)** -- A large impulsive type electromagnetic wave generated by nuclear or chemical explosions.

**EQUIPMENT, UNIT OR PIECE OF** -- An item having a complete function apart from being a component of a system.

**EQUIPMENT GROUNDING** -- Attained by the grounding conductor of the fault protection subsystem, and/or bonding to the signal reference subsystem or the structural steel elements of the building.

**EQUIPOTENTIAL PLANE** -- A grid, sheet, mass, or masses of conducting material which, when bonded together, offers a negligible impedance to current flow. (serves as signal reference subsystem for new facilities)

**FACILITY** -- A building or other structure, either fixed or transportable in nature, with its utilities, ground networks, and electrical supporting structures. All wiring, cabling as well as electrical and electronic equipments are also part of the facility.

**FACILITY GROUND SYSTEM** -- The electrically interconnected system of conductors and conductive elements that provides multiple current paths to earth. The facility ground system includes the earth electrode subsystem, lightning protection subsystem, signal reference subsystem, fault protection subsystem, as well as the building structure, equipment racks, cabinets, conduit, junction boxes, raceways, duct work, pipes, and other normally noncurrent-carrying metal elements.

**FAR FIELD** -- The region of the field of an antenna where the radiation field predominates and where the angular field distribution is essentially independent of the distance from the antenna.

**FAULT** -- An unintentional short-circuit, or partial short-circuit, (usually of a power circuit) between energized conductors or between an energized conductor and ground.

**FIRST SERVICE DISCONNECT** -- The necessary equipment (circuit breakers, switches, fuses etc.) located at the point of entrance of power conductors to a building or other structure.

**GROUND** -- The electrical connection to earth primarily through an earth electrode subsystem. This connection is extended throughout the facility via the facility ground system consisting of the signal reference subsystem, the fault protection subsystem, the lightning protection subsystem and the earth electrode subsystem.

**GROUNDING CONDUCTOR** -- (Neutral) The circuit conductor that is intentionally grounded (at first service disconnect or power source).

**GROUNDING CONDUCTOR -- (Green Wire)** A conductor used to connect equipment or the grounded circuit of a power system to the earth electrode subsystem.

**HIGHER FREQUENCY GROUND --** The interconnected metallic network (equipotential plane) intended to serve as a common reference for currents and voltages at frequencies above 30 kHz and in some cases above 300 kHz. Pulse and digital signals with rise and fall times of less than 1 microsecond are classified as higher frequency signals.

**INTERFACE --** Any electrical connection (encompassing power transfer, signaling, or control functions) between two or more equipments or systems.

**ISOKERAUNIC (or isoceraunic) --** Showing equal frequency of thunderstorms.

**ISOLATION --** Physical and electrical arrangement of the parts of an equipment, system, or facility to prevent uncontrolled electrical contact within or between the parts.

**LIGHTNING PROTECTION SUBSYSTEM --** A complete subsystem consisting of air terminals, interconnecting conductors, ground terminals, arresters and other connectors or fitting required to assure a lightning discharge will be safely conducted to earth.

**LOWER FREQUENCY GROUND --** A dedicated, single-point network intended to serve as a reference for voltages and currents, whether signal, control or power, from dc to 30 kHz and some cases to 300 kHz. Pulse and digital signals with rise and fall times greater than 1 microsecond are considered to be lower frequency signals.

**MAGNETIC FIELD --** A vector field produced by a continuous flow of charge.

**MULTIPOINT GROUND --** More than one path to ground.

**NATIONAL ELECTRICAL CODE (NEC) --** A standard governing the use of electrical wire, cable, and fixtures installed in buildings. It is sponsored by the National Fire Protection Association (NFPA-70) under the auspices of the American National Standards Institute (ANSI-C1).

**NEAR FIELD --** The region of the field immediately surrounding an antenna where the inductive and capacitive fields predominate. In this region the angular distribution of the field varies with distance from the antenna.

**NEUTRAL --** The ac power system conductor which is intentionally grounded on the supply side of the first service disconnect(ing) means. It is the low potential (white) side of a single phase ac circuit or the low potential fourth wire of a three-phase wye distribution system. The neutral (grounded conductor) provides a current return path for ac power currents whereas the grounding (or greener) conductor does not, except during fault conditions.

**PENETRATION --** The passage through a partition or wall of an equipment or enclosure by a wire, cable, or other conductive object.

**PLANE WAVE** -- An electromagnetic wave which predominates in the far field region of an antenna, and with a wavefront which is essentially in a flat plane. In free space, the characteristic impedance of a plane wave is 377 ohms.

**RACK** -- A vertical frame on which one or more units of equipment are mounted.

**RADIATION** -- The emission and propagation of electromagnetic energy through space.

**RADIATION RESISTANCE** -- The resistance which, if inserted in place of an antenna, would consume the same amount of power that is radiated by the antenna.

**RADIO FREQUENCY INTERFERENCE (RFI)** -- RFI is manmade or natural, intentional or unintentional electromagnetic propagation which results in unintentional and undesirable responses from or performance degradation or malfunction of, electronic equipment.

**REFLECTING LOSS** -- The portion of the transition loss, expressed in dB, that is due to the reflection of power at a barrier or shield. Reflection loss is determined by the magnitude of the wave impedance inside the barrier relative to the wave impedance in the propagation medium outside the barrier.

**RF-TIGHT** -- Offering a high degree of electromagnetic shielding effectiveness.

**SHIELD** -- A housing, screen, or cover which substantially reduces the coupling of electric and magnetic fields into or out of circuits or prevents the accidental contact of objects or persons with parts or components operating at hazardous voltage levels.

**SHIELDING EFFECTIVENESS** -- A measure of the reduction or attenuation in the electromagnetic field strength at a point in space caused by the insertion of a shield between the source and that point.

**SIGNAL REFERENCE SUBSYSTEM** -- A conductive sheet or cable network/mesh providing an equipotential reference for C-E equipments to minimize interference and noise.

**SIGNAL RETURN** -- A current-carrying path between a load and the signal source. It is the low side of the closed loop energy transfer circuit between a source-load pair.

**STRUCTURE** -- Any fixed or transportable building, shelter, tower, or mast that is intended to house electrical or electronic equipment or otherwise support or function as an integral element of an electronics complex.

**SUPPORTING STRUCTURES, ELECTRICAL** -- Normally nonelectrified conductive structural elements near to energized electrical conductors such that a reasonable possibility exists of accidental contact with the energized conductor. Examples are conduit and associated fittings, junction and switch boxes, cable trays, electrical/electronic equipment racks, electrical wiring cabinets, and metallic cable sheaths.

**TRANSDUCER** -- A device which converts the energy of one transmission system into the energy of another transmission system.

**THUNDERSTORM DAY** -- A local calendar day on which thunder is heard.

**UNDESIRE SIGNAL** -- Any signal which tends to produce degradation in the operation of equipments or systems.

**WAVE IMPEDANCE** -- The ratio of the electric field strength to the magnetic field strength at the point of observation.

**ZONE OF PROTECTION** -- (also known as **CONE OF PROTECTION**) That space that is below and adjacent to a lightning protection subsystem that is substantially immune to direct lightning discharges.

APPENDIX B

SUPPLEMENTAL BIBLIOGRAPHY

PART I. SUBJECT CROSS REFERENCE.

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(Project SLHC-4191)

**Preparing Activity:**

Air Force - 90

**Review Activities:**

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